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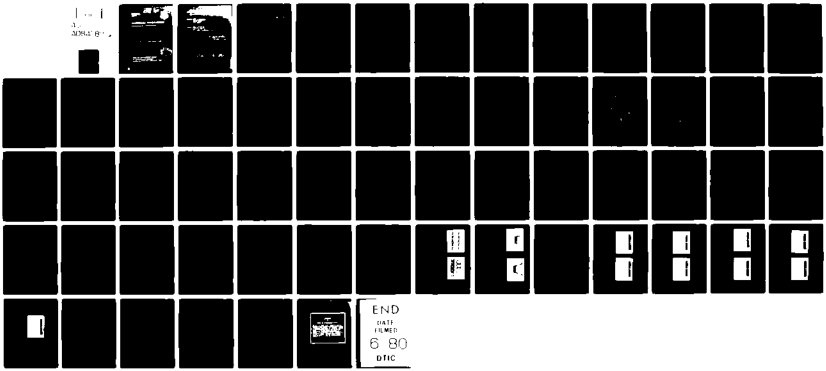

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20. ABSTRACT (Continue on reverse side if necessary and identify by block number)

This report describes a research effort intended to determine the feasibility of combining high power solid state amplifier and electronically tunable resonator technology. This new technology would be applicable to the design and fabrication of high power electronically tunable amplifiers capable of frequency agile operation in the UHF range.

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JOB

The primary goal of the effort was to determine if the RF power from twelve 100 watt solid state amplifier modules could be combined in the electric field of an electronically and/or mechanically tuned cavity resonator, to produce/kilowatt of continuous output power. Testing of the experimental model showed that efficient RF power combining could be obtained in a mechanically tuned cavity where the length of the coaxial inner element was varied to change the frequency. Summing with the electronic tuning technique (PIN diode switched loops) was inefficient and distortion of the RF field by the tuning loops within the cavity precludes tuning and summing at the same time.

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EVALUATION

The objective of this research effort was to determine the feasibility of dividing and combining RF power from solid state radios and amplifiers within a mechanically or electronically tuned resonator. This effort provided the initial research in support of the "EMC Technology for Frequency Hopping Systems" program outlined in RADC technical program objectives (TPO) R4C - Electromagnetic Compatibility. The specific problem addressed was the reduction of unwanted spurious and broadband noise outputs from high power frequency agile emitters. Results of the effort showed that it was possible to efficiently combine the outputs of 12 each 100 watt amplifiers in a mechanically tuned cavity. However, in the electronically tuned case, the tuning technique chosen distorted the electric field within the cavity and produced unsymmetrical coupling to and from the amplifiers. Recommendations and conclusions derived from this research will be used as inputs to future exploratory development efforts addressing the EMC problems associated with high power, frequency agile operation.

Thomas E. Baustert

THOMAS E. BAUSTERT
PROJECT ENGINEER

1.0 INTRODUCTION

This report describes the work that was performed by the Magnavox Government and Industrial Electronics Company under Rome Air Force Development Center Contract F30602-78-C-0152 to develop a one kilowatt Solid State Electronically Tuned Amplifier (ETA) during the period April 1978 to April 1979.

1.1 Objectives. The objectives of the program were to conduct, design and produce an experimental hardware prototype RF amplifier demonstrating qualities necessary for low noise, frequency agile operation in the UHF military communications band. The goal of the program was to show that the RF power of several modular amplifiers could be summed efficiently within the electric field of a cavity resonator and that the resonator could be tuned over the frequency range 336 to 400 MHz by both electronic and manual means.

1.2 Results of Investigation. It was found that the energy from twelve 100 watt RF amplifier modules could be summed directly in the electric field of a cavity resonator to produce one kilowatt continuous output power. The investigation also showed that electronic tuning of the cavity resonator is feasible only when RF summing does not take place within the electric field of the resonator.

1.3 Summary of Accomplishments. Under Contract F30602-78-C-0152 an amplifier was developed which produced 1 KW of continuous output power over the frequency range 336 to 400 MHz. The unit is packaged in a 1-1/2 ATR case exclusive of power supply and a fan for forced air cooling. Electronic tuning was not incorporated in the amplifier but was investigated using a separate resonator similar to that used in the amplifier. Electronic frequency switching was demonstrated at a power level of 350 watts using a single input port to the cavity.

2.0 DESCRIPTION OF APPROACH

The Electronically Tuned Amplifier is comprised of a tuned input resonator, twelve 100 watt amplifier modules, an output resonator, and electronically controlled attenuator as shown in Figure 2-1. The 100 watt

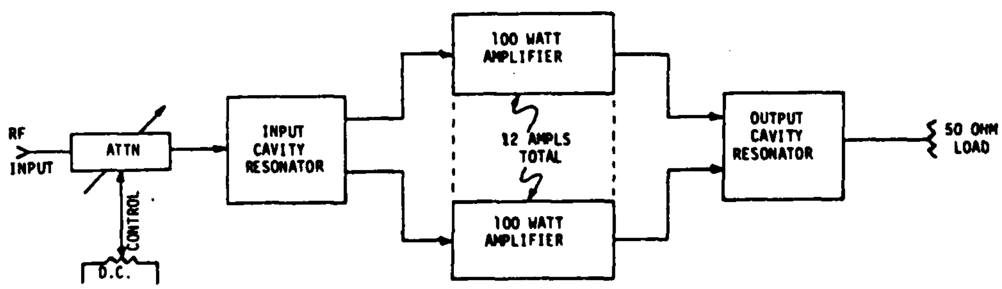


FIGURE 2-1. 1 KW RF AMPLIFIER BLOCK DIAGRAM

modules were designed for mounting directly on the resonator assembly with summing of RF power taking place within the cavity by means of loop coupling from the output transistor collector circuit. The input cavity divides the RF input into twelve channels to drive the amplifier modules. An electronically controlled attenuator is provided at the RF input for operating the system at various power levels. Originally, it was planned to use a microprocessor controlled closed loop gain leveling scheme in conjunction with electronic tuning of the cavity resonators. It was found that leveling of amplifier power with electronic tuning was not feasible due to variations between the load impedances at each of the twelve cavity resonator input ports under conditions of electronic tuning. The ETA was therefore implemented as a purely manually tuned system with tuning accomplished by mechanically varying the length of the coaxial arm of the resonator.

2.1 Modular Amplifiers. The ETA system was designed for 1000 watts maximum continuous power output. It was calculated that the total power loss in the output resonator would not exceed 0.8 dB. This required 1200 watts from the solid state amplifiers, or twelve 100 watt units. A Communications Transistor Corporation C2M100-28A was selected for use as the output devices. This choice was based on gain considerations and the fact that a single ended, low impedance design was required. The C2M100-28A also offered resonable efficiency and the best heat transfer characteristic available.

The minimum input signal available to the system was specified as 63 watts (100 watts - 2 dB). An input resonator was used for power division to the 12 amplifiers. It was calculated to have an 0.8 dB loss. In addition, each amplifier plus the system itself has a voltage variable attenuator. Minimum insertion loss was established to be 1 dB for each attenuator. This results in a total loss in input signal of 2.8 dB (1 + 1 + .8 = 2.8). The power available is then:

$$63 \text{ watts} - 2.8 \text{ dB} = 33.1 \text{ watts}$$

or

$$\frac{33.1}{12} = 2.76 \text{ watts per amplifier}$$

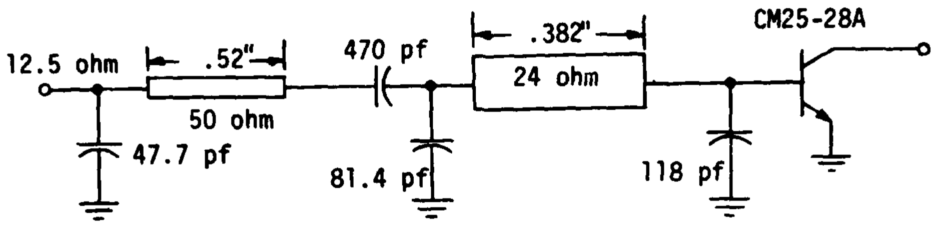
The minimum gain required per amplifier is therefore:

$$10 \log_{10}\left(\frac{100}{2.76}\right) = 15.6 \text{ dB}$$

Therefore, a two stage amplifier is required. A CTC CM25-28A transistor was slected for the driver amplifier because of its high gain characteristics. Both the driver and output transistors are rated to operate into a 3:1 VSWR without damage. Nominal gain expected was 9.5 dB in the driver stage and 8 dB in the power output stage at 400 MHz. This results in a typical overall gain of 17.5 dB with a 100 Watt power output level. Increased gain at 336 MHz plus a ± 1.5 dB variation between units can result in a gain spread of 16 to 21 dB. All excess input power was to be dissipated in the system attenuator.

The amplifiers were designed with copper heat sinks to secure best heat conduction away from the transistor junction. PC board material is 0.0315" thick fiberglass with 1 oz. copper both sides. Plated through holes are used to join the top and bottom ground planes. Eyelets are used in several critical areas where high rf circulating currents exist. Transmission lines (microstrip) are used for series and shunt inductances. Twenty-four ohm transmission lines are used for the base and collector connections. Chip capacitors (ATC 100 series) are used in the rf impedance matching cirucits except at the high current points in the output stage. Here, ATC 175 series capacitors are used because of their lower equivalent series resistance. All impedance matching was done using "COMPACT"* computer aided design techniques.

The final amplifier input matching circuit as optimized by COMPACT is shown in Figure 2.2.



TRANSISTOR INPUT IMPEDANCE

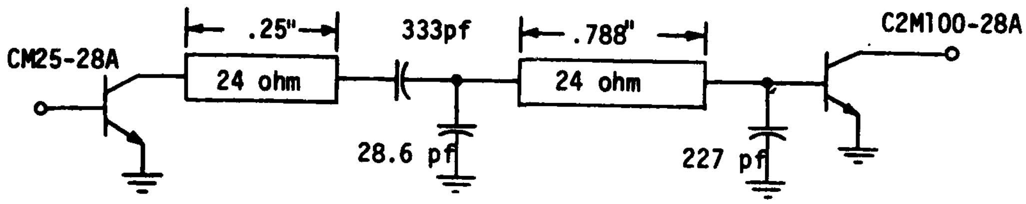
f-MHz	Zi-OHMS	
336	1.9 + j1.1	estimated
356	2.09 + j1.3	estimated
377	2.23 + j1.6	estimated
400	2.4 + j1.9	Measured

FIGURE 2-2 AMPLIFIER INPUT CIRCUIT

The computed VSWR to the 12.5 ohm input was 1.06:1 or better between 336 and 400 MHz. A 4:1 impedance transformer is used to provide a match between the 50 ohm source and the circuit input. This was made with a 2 inch piece of 25 ohm microcoax cable twisted into 3 turns. Experiments with the breadboard circuit resulted in final capacitor values of 33 pf in place of the 47.7, two 24 pf's in parallel in place of the 81.4 and two 36 pf's in place of the 118.

The final interstage matching circuit between the CM25-28A driver and the C2M100-28A output stage as optimized by COMPACT is shown in Figure 2-3.

*COMPACT Engineering, Inc.
Los Altos, CA



CM25-28A

OUTPUT IMPEDANCE (CALCULATED)

f-MHz	Z1-OHMS
300	9.25 - j8.01
325	8.47 - j 7.77
350	7.76 - j7.16
375	7.12 - j7.16
400	6.54 - j6.81
425	6.02 - j6.45

C2M100-28A

INPUT IMPEDANCE CTC DATA

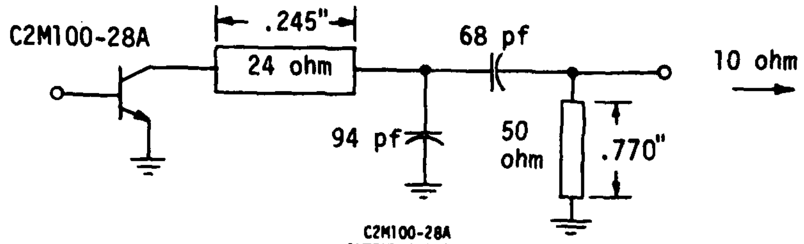
f-MHz	Z1-OHMS
300	.52 + j1.4
325	.6 + j1.48
350	.79 + j1.54
375	.94 + j1.54
400	1.09 + j1.49
425	1.19 + j1.42

FIGURE 2-3 AMPLIFIER INTERSTAGE CIRCUIT

The computer optimization was done for the best match to the driver. Experiments with the breadboard circuit resulted in the use of a 300 pf capacitor in place of the 333, two 12 pf's in parallel in place of the 28.6 and two 36 pf's in parallel in place of the 227.

The biggest problem area in the design of the 100 watt amplifier was in achieving a satisfactory output matching circuit. One of the objectives of this program was to combine a number of high power transistor amplifiers at a very low impedance which approached the output impedance of the transistor itself. However, interconnection considerations dictated the use of a real impedance high enough to permit fabrication of a transmission line to conduct the rf power from the transistor collector to the resonator input loop. A five ohm impedance proved to be the lowest practical and this value was used in the initial design. A major problem was discovered in the bread board circuit that made this approach unsatisfactory. No capacitors could be found that could withstand the very high circulating rf currents in the matching circuits at 100 W cw power level.

This problem was overcome by redesigning the output stage to match into a 10 ohm load. The 10 ohm output matching circuit as optimized by COMPACT is shown in Figure 2-4.



C2M100-28A
OUTPUT IMPEDANCE
(CALCULATED)

f-MHz	Zo-OHMS
300	1.92 + j0.54
325	1.81 + j0.67
350	1.71 + j0.8
375	1.62 + j0.94
400	1.53 +j1.09
425	1.44 +j1.24

FIGURE 2-4 AMPLIFIER OUTPUT CIRCUIT (10 OHM)

The computed VSWR to the 10 ohm output load was better than 1.32:1 between 325 and 400 MHz. Experiments with the breadboard circuit resulted in the use of two 56 pf capacitors in parallel and one 56 pf in series for best operation. These were ATC 175 series for minimum loss.

Twelve of these amplifier modules were built and tested. The amplifier gain varied from a low of 18.9 dB and a high of 22.6 dB. The average efficiency of the output stages was 53.2% with a unit low of 50.1%.

Two major problems became apparent when several of these modules were operated on the cavity resonators. The first was determined to be due to a lack of isolation between amplifier modules making it impossible to balance units by varying the individual input powers. An amplifier could be made to draw considerable current in the output stage even with no rf input signal.

The second problem was due to the design of the coupling loops on each amplifier's output to the resonant cavity. It was revealed by increasing difficulty in operation as more amplifiers were connected into the system. It was originally believed that the impedance seen by each amplifier was independent of the number of amplifiers, i.e., twelve amplifiers with 10 ohms output impedance would each want to drive a 10 ohm loop as measured by itself. This did not prove to be the case, but rather, each loop wanted to look like 0.833 ohms $(\frac{10}{12} = 0.833)$. Since this impedance was too low to implement, it was decided that the amplifiers should be modified to drive 50 ohm loads. Then the required loop impedance is 4.17 ohms $(\frac{50}{12} = 4.17)$. This is easily attained. The final amplifier output circuit as optimized by COMPACT is shown in Figure 2-5.

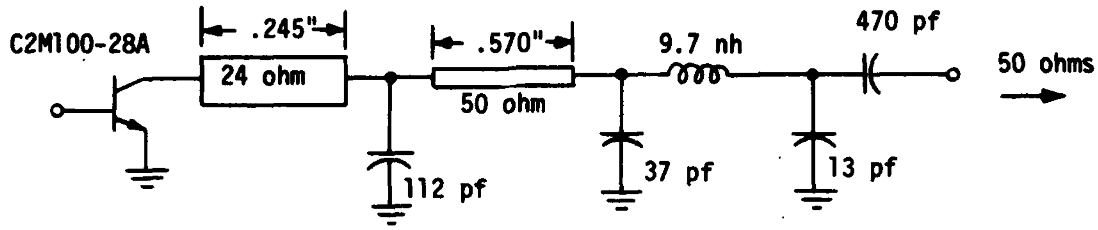
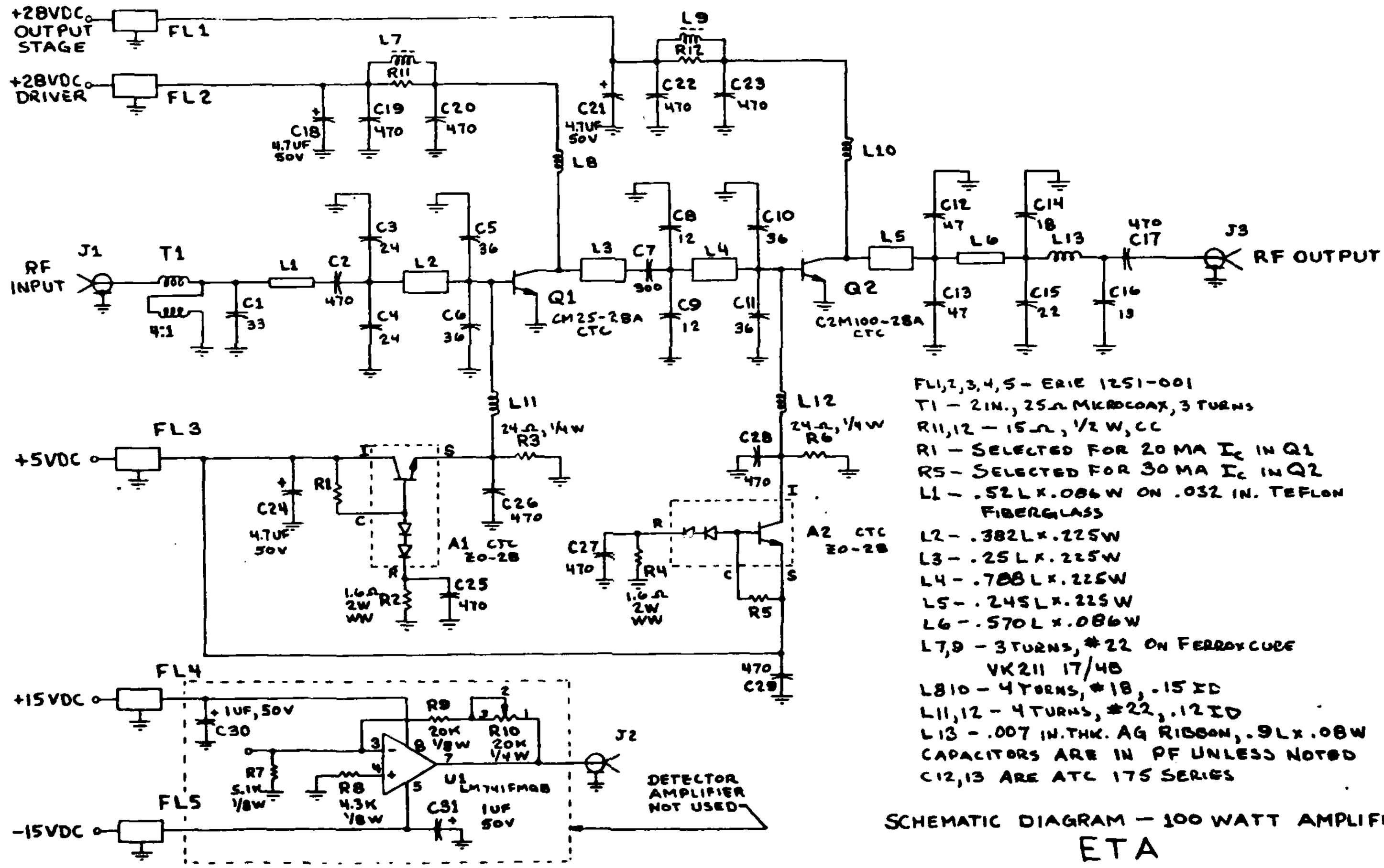


FIGURE 2-5. AMPLIFIER OUTPUT CIRCUIT (50 OHMS)

This circuit was designed with a primary consideration being a minimum of changes to the PC board of the 10 ohm circuit. The VSWR was computed to be better than 1.3:1 between 325 and 425 MHz. The 9.7 nh inductor is formed from a silver ribbon 0.007 in. thick by 0.08 in. wide by approximately 0.9 in. long. The 112 pf capacitance is two 47 pf ATC 175 series capacitors in parallel. The 37 pf is an 18 and 22 pf ATC 100 series capacitors in parallel. The amplifier gain varied from 18.5 to 21.9 dB. The average efficiency of the output stages was 54.4% with a unit low of 48.2%. Figure 2-6 is the schematic diagram of the amplifier module as used in the ETA system.

When the first amplifier (10 ohms output impedance) was attached to the output cavity resonator a low power oscillation appeared at 420 MHz. It was only present when input power was applied. This same problem was present when the amplifier was connected to a tunable bandpass filter on its output. The only effective solution to this problem was in providing a very short ground path between the two emitter (ground) leads and the isolated mounting flange. This was accomplished by soldering a short length of heavy buss wire between the mounting flange and emitter leads and up against the transistor body. This was done on both sides. Communications Transistor Corporation was contacted but could offer no explanation for the problem or solution.

2.2 Cavity Resonators. Two coaxial cavities, input and output, electronically tuned form the basis of the electronically tuned amplifier. Mechanical tuning is also incorporated which allows the ETA to be tuned manually over the frequency range of 336 MHz to 400 MHz. Electronic tuning is accomplished by short circuiting inductive tuning loops (3), placed at the short circuited end of the cavity resonators, allowing frequency changes in 16 MHz steps. The inductive tuning loops are short circuited by applying an appropriate forward bias to PIN diodes that are placed in series with each inductive tuning loop. The input cavity resonator performs the function of an



FL1,2,3,4,5 – ERIE 1251-001
T1 – 2IN., 25Ω MICROCOAX, 3 TURNS
R11,12 – 15Ω, 1/2 W, CC
R1 – SELECTED FOR 20 MA I_C IN Q1
R5 – SELECTED FOR 30 MA I_C IN Q2
L1 – .52 L x .086 W ON .032 IN. TEFLON FIBERGLASS
L2 – .382 L x .225 W
L3 – .25 L x .225 W
L4 – .788 L x .225 W
L5 – .245 L x .225 W
L6 – .570 L x .086 W
L7,9 – 3 TURNS, #22 ON FERROXCUBE VK211 17/4B
L8,10 – 4 TURNS, #18, .15 ID
L11,12 – 4 TURNS, #22, .12 ID
L13 – .007 IN. THK. AG RIBBON, .9 L x .08 W
CAPACITORS ARE IN PF UNLESS NOTED
C12,13 ARE ATC 175 SERIES

SCHEMATIC DIAGRAM — 100 WATT AMPLIFIER
ETA
FIGURE 2-6

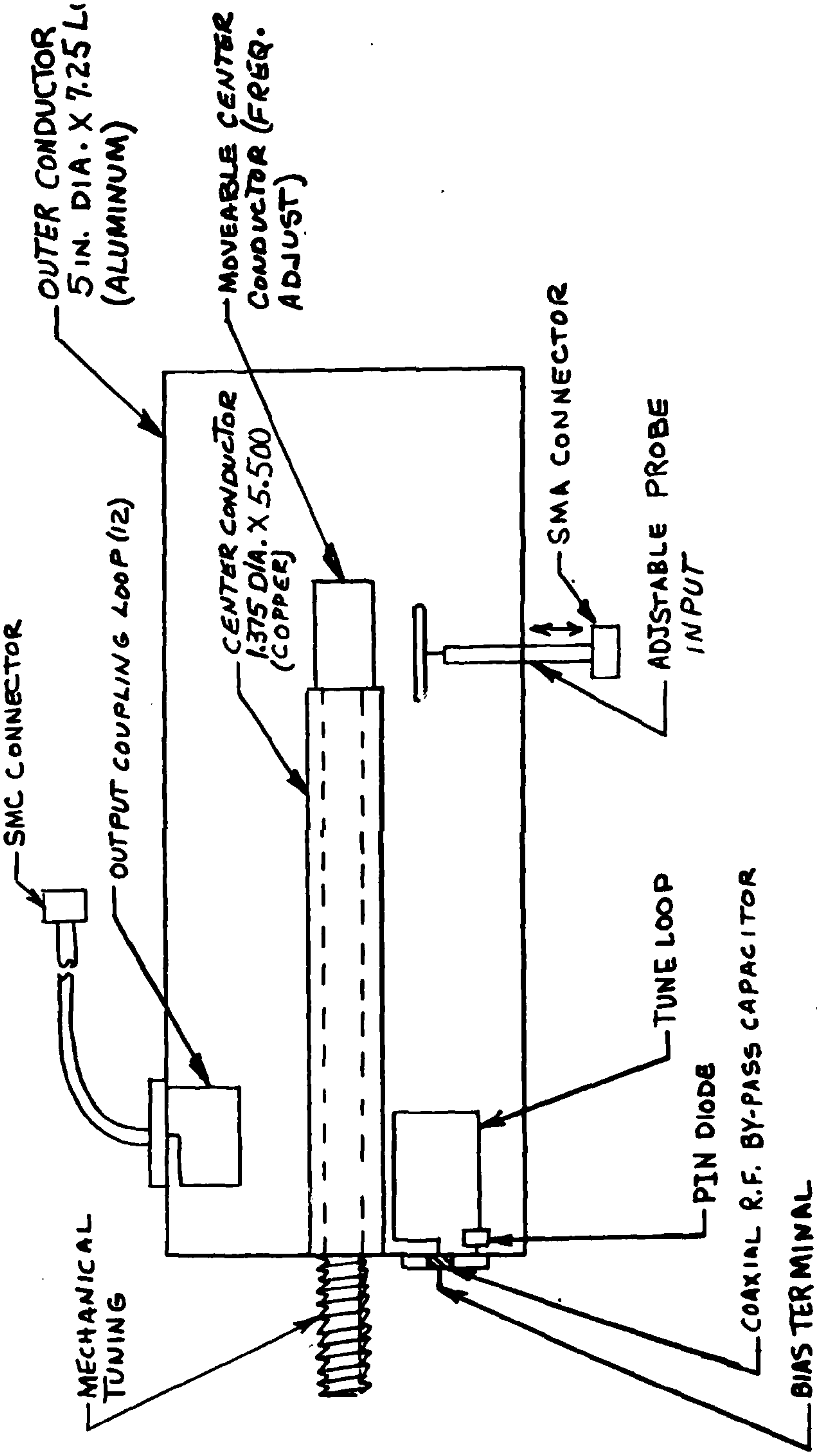
in-phase equal 12 way RF power divider that feeds 12 modular 100 watt RF power amplifiers. The RF outputs (12) are combined in the output cavity resonator. Power is coupled in and out of the cavity resonators by utilizing capacitive probes inserted into the cavity near the open circuited end of the cavity resonators.

2.2.1 Input Cavity Resonator. The input coaxial cavity resonator performs the function of an in-phase 12 way RF power divider and a single pole resonator filter. The basic design objectives for the developmental input cavity resonator were:

Frequency Range	336 to 400 MHz
Electronic Tuning	352 to 400 MHz (4 discrete steps 16 MHz ea.)
Mechanical Tuning	336 to 400 MHz (continuous)
RF Input Power	200 watts CW Max
Loaded Q (Q_L)	125
Unloaded Q (Q_u)	2500 Min.

The mechanical design of the ETA dictated the basic cavity resonator size. An outer diameter of 5 inches and length of 7.25 inches was selected. These dimensions allowed the mounting of the electronic control modules and output coupling loops around the outer pereripery of the coaxial resonator. Figure 2-7 is a simplified cross section of the input cavity resonator. The coaxial cavity resonator center conductor diameter was selected for optimum unloaded Q. The theoretical unloaded Q for the input resonator is 9967 at 352 MHz. The measured Q of the input resonator is approximately 65% of the theoretical unloaded Q. Most of the losses were due to the center conductor. The unloaded Q was still enough to insure very low insertion loss. The input to the cavity resonator is a simple adjustable capacitive probe inserted near the high impedance plane of the coaxial cavity resonator. The output coupling utilizes 12 equal spaced inductive loops, that can be rotated, located around the outer pererphery of the cavity resonator and located near the short circuited end of the resonator. The output coupling loop size was selected to present a nominal 50 ohm output impedance when each of 12 loops are terminated into 50 ohm loads. The output impedance of each inductive loop is 1/12 of 50 ohms or 4.17 ohms when all of the other 11 output loops are terminated into 50 ohm loads. The measured loaded Q varied from 57 at 400 MHz to 71.7 at 352 MHz. The maximum measured insertion loss was 1.4 db over the frequency band of 352 MHz to 400 MHz with the tune loops installed. The electronic tuning was not incorporated in the final version of the ETA due to two main problems experienced in the early phases of the development program. The main problem was the field distortion within the cavity near the output coupling loops caused by the tune loops when they were short circuited. This field distortion caused varying impedance changes reflected into the output coupling loops and a variation in the power output coupling from each coupling loop. The other problem was the voltage breakdown or rectification that occurred in the PIN diodes when they were reversed bias and at the same time subjected to RF power levels in excess of 350 watts CW. The mechanical tuning was incorporated in final ETA version.

[1] Terman, Radio Engineers Handbook, McGraw-Hill 1943



ETA INPUT CAVITY RESONATOR

FIGURE 2-7

The electronic tuning was implemented, along with the mechanical tuning, by inserting three appropriate inductive tuning loops located at the short circuited end of the coaxial cavity resonator. The tune loops were integrated with a PIN diode which allowed the loop to be short circuited or "open" with forward or reverse bias. This technique was used to obtain the four discrete frequency steps of 16 MHz starting at 352 MHz. The tune loops were installed and tested after the output coupling loops had been characterized. The tune loop dimensions were initially calculated from the following equation:[2]

$$X_M = \frac{AufCos\theta_1}{r}$$

where:

X_M = Mutual Reactance in ohms
A = Area of tune loop in meters^2
u = $4\pi \times 10^{-7}$
r = Radial distance from center of center conductor to center loop in meters
f = Frequency in Hz
θ_1= Electrical degrees from the short circuit to center of loop

The above equation does not take into account the self inductance of the loop and the capacitive reactance which is reflected into the coaxial cavity resonator when the tune loop coupling is varied. The tune loop assembly was designed in such a manner that the coupling into the cavity resonator could be varied. The final tune loop dimensions were obtained empirically by fabricating different inductive loop sizes in a fixture which contained the PIN diode and the RF by-pass capacitor. This fixture could then be placed into the cavity resonator and characterized. Figure 2-8 is a typical curve showing the tune loop size versus frequency shift when the cavity resonator is initially tuned to 352 MHz with no tune loops installed. Figure No. 2-9 is a typical curve showing the total frequency shift caused by three tune loops placed inside the cavity resonator. The Freq shift curves shown in Figure 2-8 and 2-9 were obtained by initially tuning the cavity resonator to 352 MHz with no tune loops installed. When the tune loops are installed, under reverse bias conditions, the resonant frequency of the cavity is shifted lower in frequency as shown. When the PIN diode is forward biased the cavity resonant frequency is shifted higher in frequency as shown. Figure 2-10 is a typical curve showing the measured self resonant frequency of various loop configurations. The loops were configured as shown in Figure 2-8 and 2-9. This data shows the importance and limitations of selecting a particular loop configuration. These measured data were utilized in designing the tune loops for the ETA.

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2 Terman, Very High-Frequency Techniques BPP773-774, Boston Technical Publishers Inc. 1965

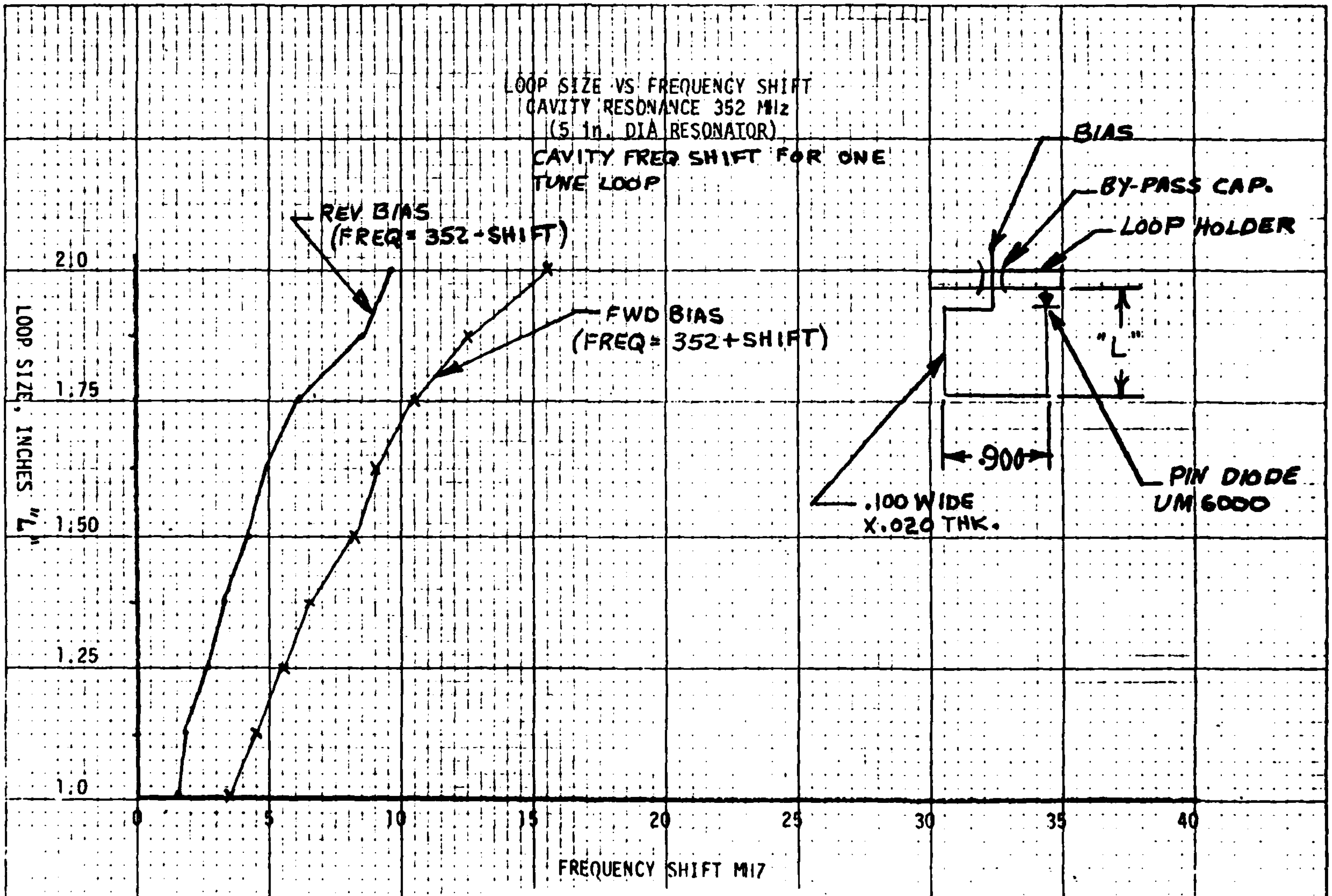
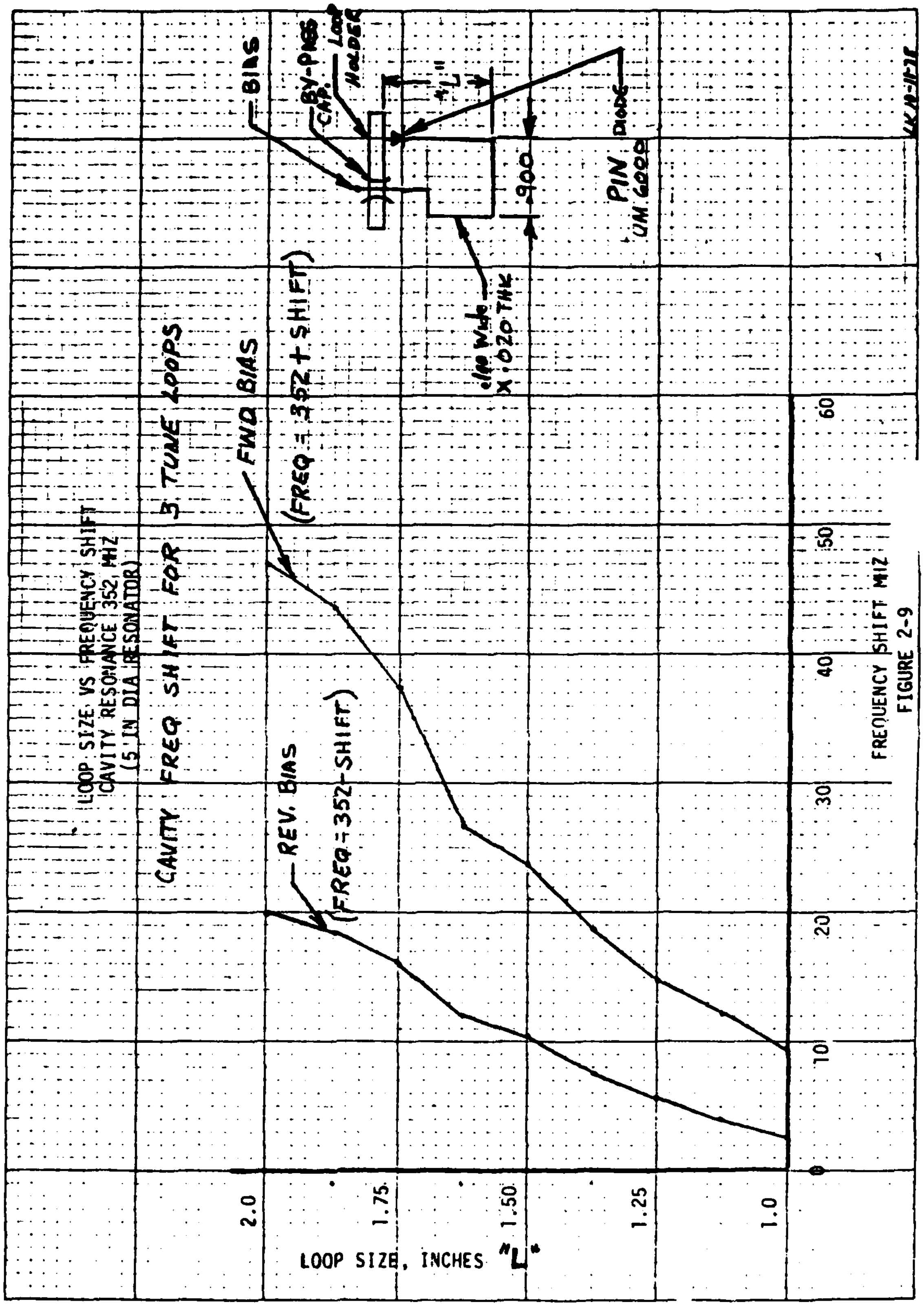

LOOP SIZE VS FREQUENCY SHIFT
CAVITY RESONANCE 352 MHz
(5 in. DIA RESONATOR)
CAVITY FREQ SHIFT FOR ONE TUNE LOOP
REV BIAS
(FREQ = 352 - SHIFT)
FWD BIAS
(FREQ = 352 + SHIFT)
BIAS
BY-PASS CAP.
LOOP HOLDER
"L"
.900
PIN DIODE
UM 6000
.100 WIDE
X .020 THK.
LOOP SIZE, INCHES "L"
2.0
1.75
1.50
1.25
1.0
0
5
10
15
20
25
30
35
40
FREQUENCY SHIFT MHz
LK 10-11-78

FIGURE 2-8



LOOP SIZE VS FREQUENCY SHIFT
CAVITY RESONANCE 352 MHZ
(5 IN DIA RESONATOR)
CAVITY FREQ SHIFT FOR 3 TUNE LOOPS
FWD BIAS
(FREQ = 352 + SHIFT)
REV. BIAS
(FREQ = 352 - SHIFT)
BIAS
BY-PASS CAP.
LOOP HOLDER
"L"
.900
PIN DIODE
UM 6000
.100 Wide X .020 THK
2.0
1.75
1.50
1.25
1.0
LOOP SIZE, INCHES "L"
0
10
20
30
40
50
60
FREQUENCY SHIFT MHZ

FIGURE 2-9

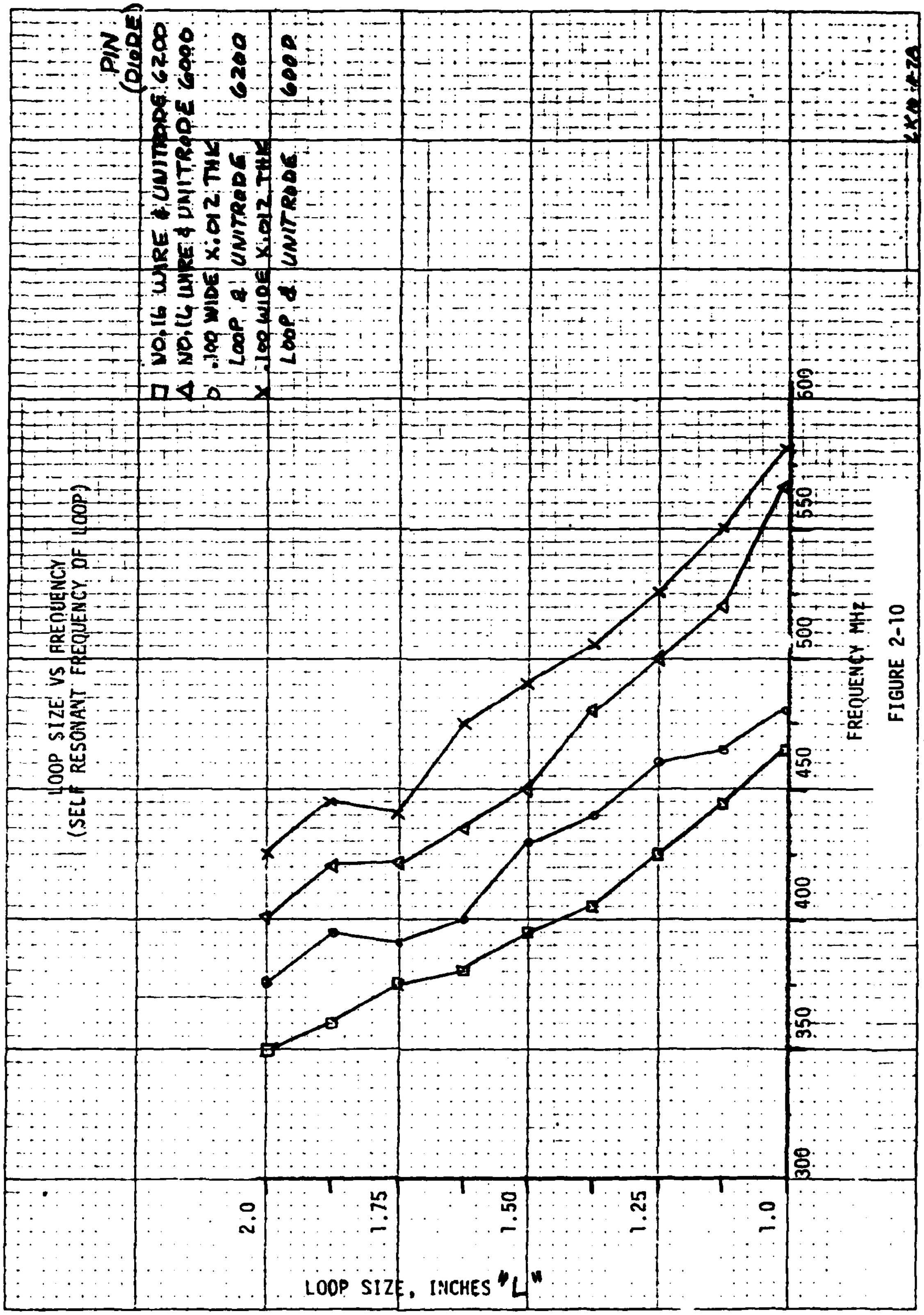
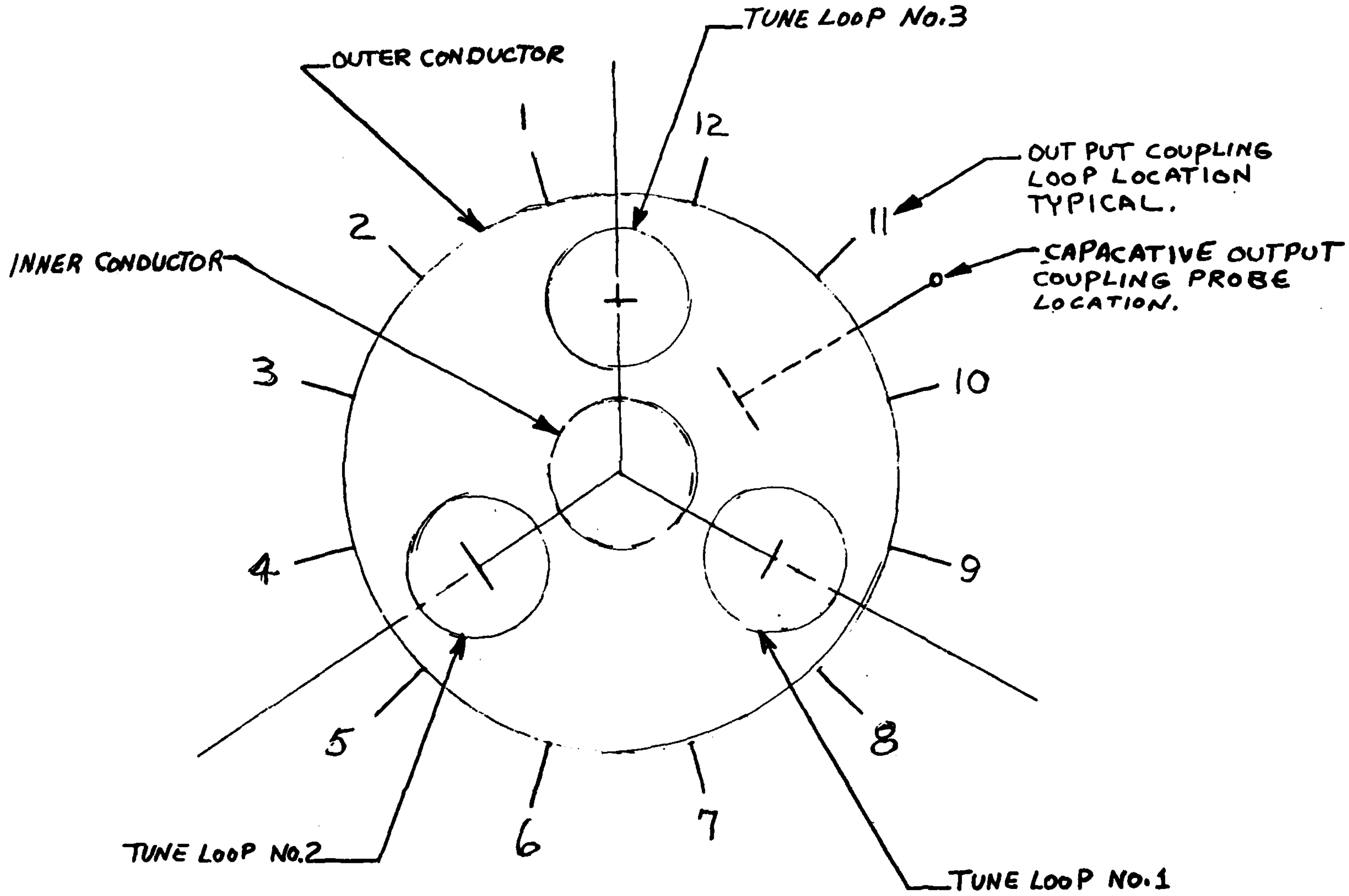

LOOP SIZE VS FREQUENCY
(SELF RESONANT FREQUENCY OF LOOP)
PIN (DIODE)
□ NO.16 WIRE & UNITRODE 6200
△ NO.16 WIRE & UNITRODE 6000
○ .100 WIDE X .012 THK LOOP & UNITRODE 6200
X .100 WIDE X .012 THK LOOP & UNITRODE 6000
300
350
400
450
500
550
600
FREQUENCY MHz
2.0
1.75
1.50
1.25
1.0
LOOP SIZE, INCHES "L"

FIGURE 2-10

As mentioned earlier, the field distribution within the resonant cavity, at the short circuited end, was investigated in the early phase of the development program when the output amplifier modules were integrated with the input cavity resonator. During the initial testing it was observed that the individual amplifier modules were not combining the power equally when the tune loops were installed. A series of RF measurements were made to determine the field distribution within the cavity resonator. The measurement was performed by applying 0 dbm (1 millwatt into 50 ohms) to the input port of the cavity resonator and measuring the output power at each one of the 12 output coupling loops when the tune loops were energized. Figure 2-11 shows the mechanical position of the tune loops, output coupling loops and the capacitive input probe. The positions of the loops are also shown in Figure 2-12 thru 2-15 which show the measured power distribution at each of the 12 output coupling loops. The curves show the actual power distribution at each output coupling loop when the tune loops are all reverse biased and when each succeeding tune loop is forward biased. The maximum power variation at each frequency is expressed in db as shown in Figure 2-12 thru 2-15. Figure 2-16 is a typical power distribution within the cavity resonator when there are no tune loops in the cavity resonator. This curve shows the even power distribution as measured at each of the 12 output coupling loops. The variation in the power distribution within the input cavity resonator caused the input power to vary to each amplifier module and in turn caused the power output of each of the 12 amplifier to vary.

The PIN diodes used to short circuit the inductive tune loops were unable to withstand RF powers in excess of 100 watts CW. The worst condition is where voltage breakdown or rectification occurs when the PIN diodes are in the reverse bias state and subjected to RF power at the same time. The voltage breakdown or rectification occurs when the tune loop and diode are subjected to RF voltage in excess of the reverse breakdown voltage of the PIN diode. Three tune loops which were designed to shift the frequency in 16 MHz steps were fabricated using Unitrode PIN diodes (UM7301) and installed into the input cavity resonator. The tune loop sizes were 2 in. x .900 in., 1.95 in. x .90 in. and 1.650 in. x .90 in. Reverse breakdown or rectification was observed to be above a point where the reverse biased diode (-50 V dc) began to conduct current greater than 1 micro ampere. The cavity resonator was tuned to 352 MHz, reverse bias was applied to the PIN diodes and RF power (CW) was applied. The point where current began to flow through the PIN diode at 1 microampere, the RF input power was 95 watts CW. The D.C. reverse breakdown of the diode was measured to be between 420 and 600 volts. The breakdown occurred in the 2.0 in. x .900 in. tune loop which has the largest cross sectional area exposed within the cavity resonator. By reducing the cross sectional area of the tune loop the breakdown of the PIN diode is reduced. Tests were performed with tune loops. 1 in. x .800 in. at 352 MHz and were able to withstand 350 watts CW. However with the reduced cross sectional size the frequency shift was correspondingly lowered from 16 MHz for the 2 in. x .90 in. loop to 3 MHz for the 1 in. x .80 in. loop. By decreasing the cross sectional area of tune loops, more loops are required to obtain the equivalent frequency shifts obtainable from one large loop.



INPUT CAVITY RESONATOR - END VIEW (SHORT CIRC. END)

FIGURE 2-11

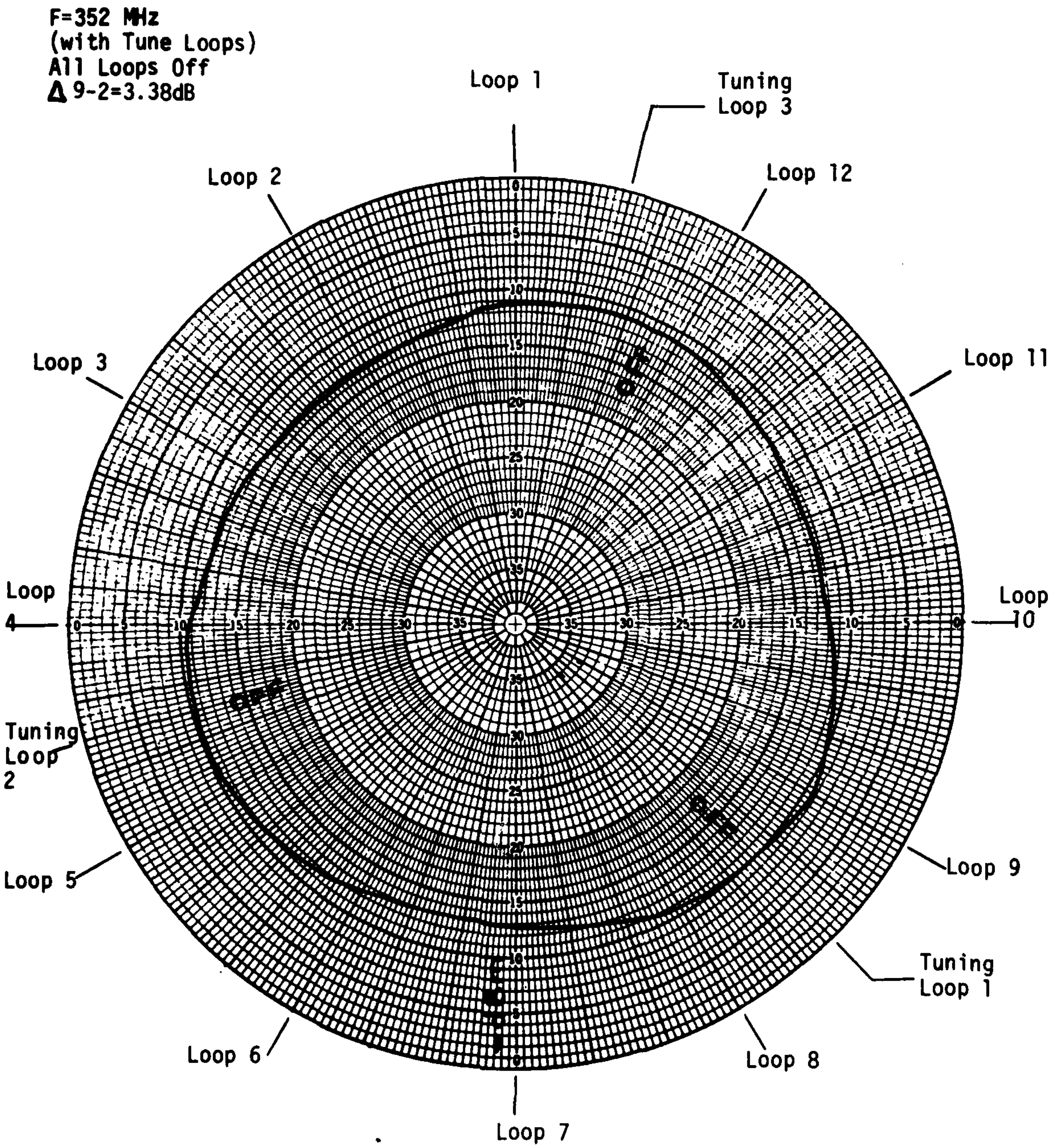


FIGURE 2-12

POWER OUTPUT VS LOOP ORIENTATION

INPUT CAVITY

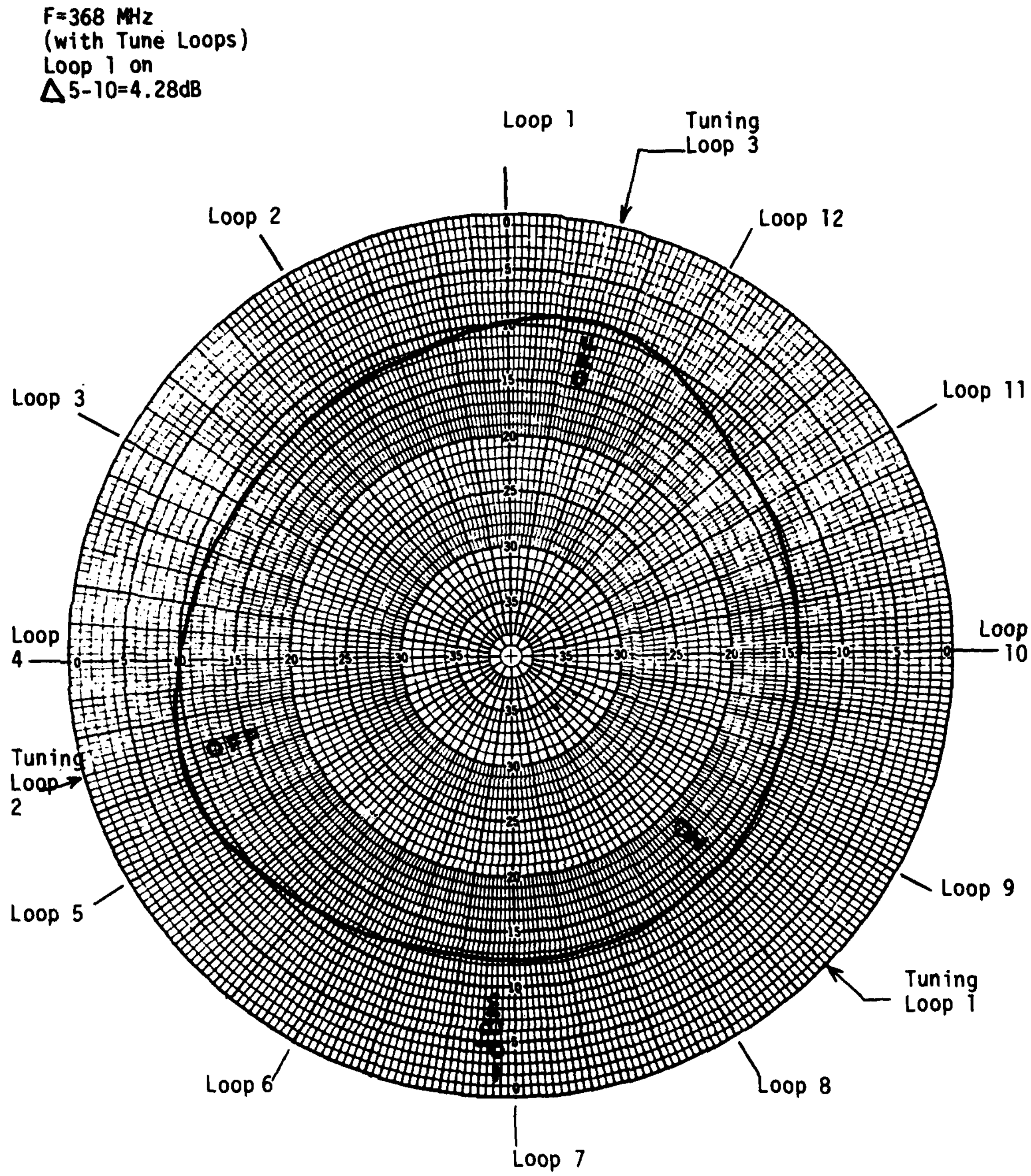


FIGURE 2-13

POWER OUTPUT VS LOOP ORIENTATION

INPUT CAVITY

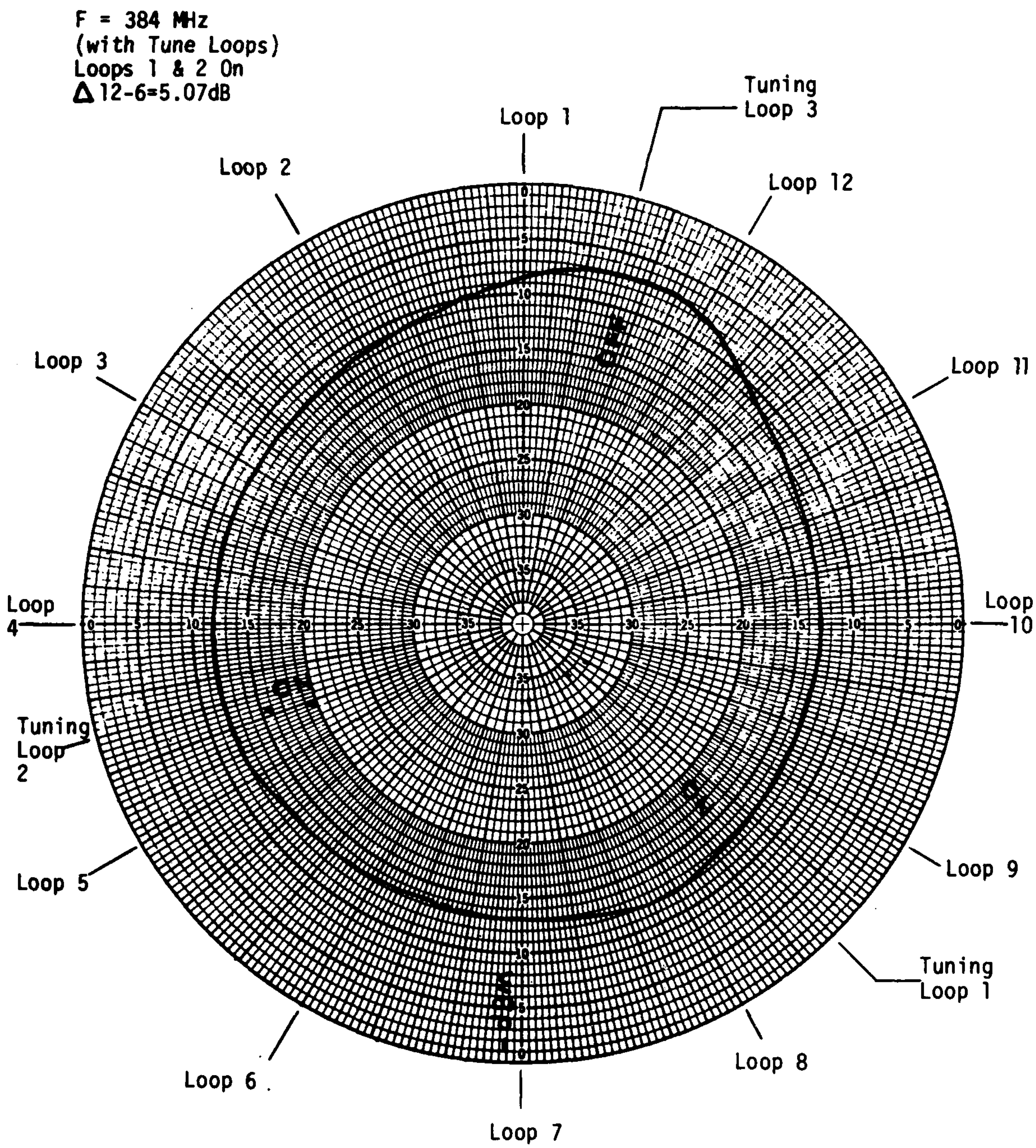


FIGURE 2-14

POWER OUTPUT VS LOOP ORIENTATION

INPUT CAVITY

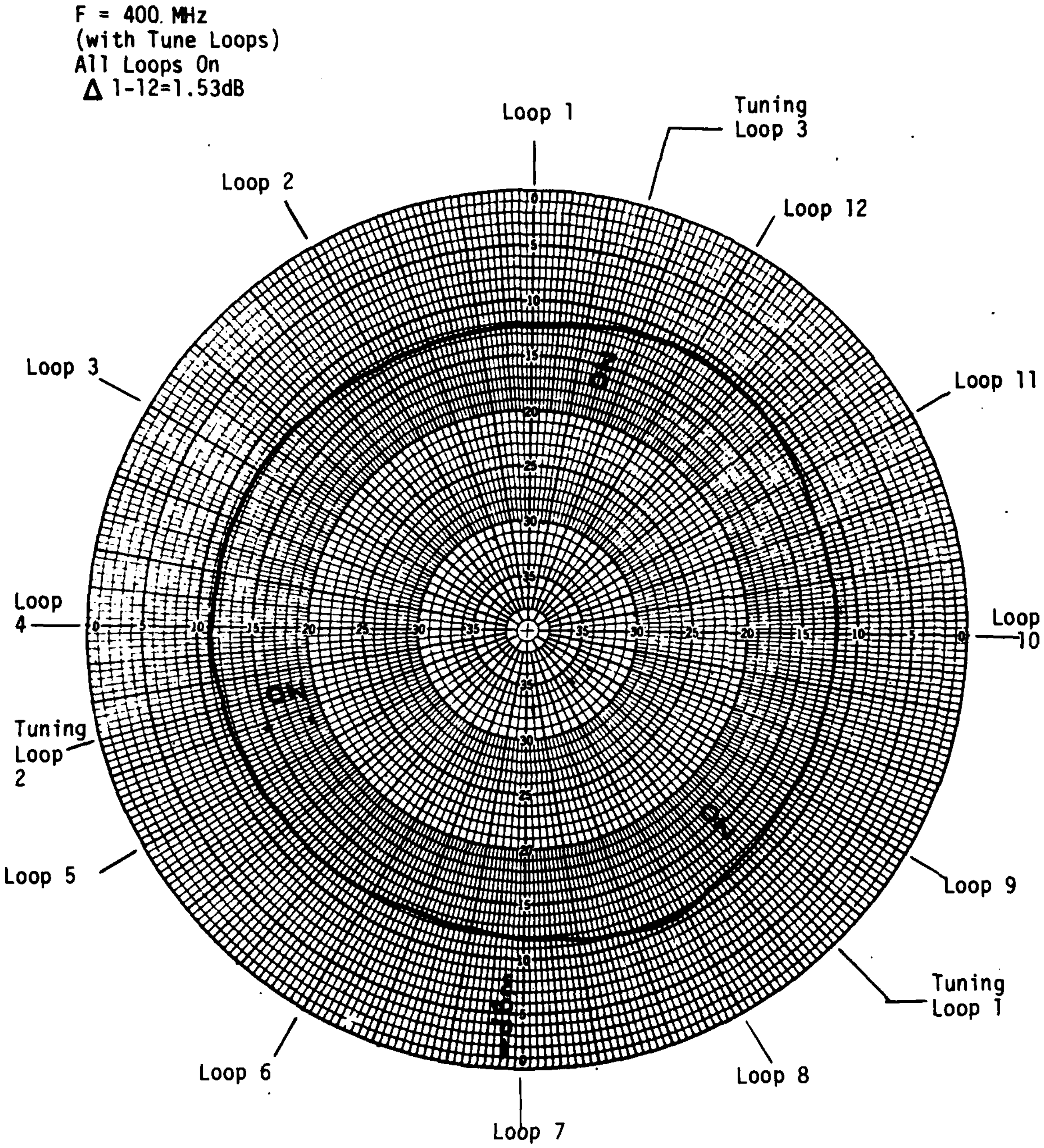


FIGURE 2-15

POWER OUTPUT VS LOOP ORIENTATION

INPUT CAVITY

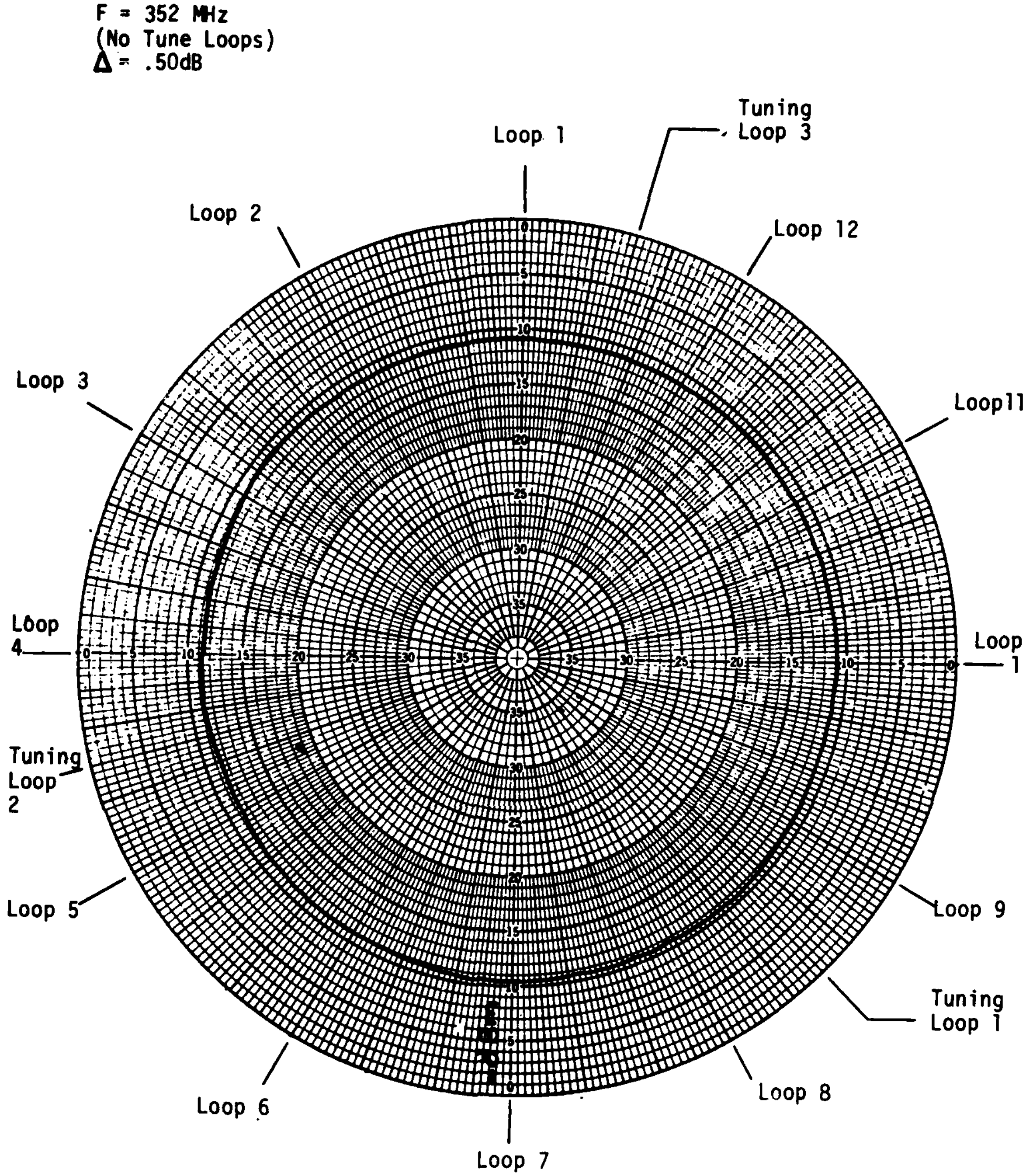


FIGURE 2-16

POWER OUTPUT VS LOOP ORIENTATION

INPUT CAVITY

The uneven power combining was degraded further due to the variation in the field distribution of the output cavity resonator near the coupling loops. This variation was caused by the tune loops which distorted the field near the short circuited end of the cavity resonator.

2.2.2 Output Cavity Resonator. The output cavity resonator performs the function of an in-phase 12 way power combiner and a single pole filter. The output cavity resonator design objectives were the same as for the input cavity resonator except the power handling requirement was 1000 watts CW. The basic construction was identical to the input resonator (see Figure 2-7) except the input coupling loops connect directly to each of the 12 100 watt amplifier modules, the center conductor is 4.5 inches long and the outer conductor is 8.25 inches long.

The input coupling loops were initially designed to be 10 ohms impedance (output impedance of amplifier modules). The output impedance of the amplifier modules was later redesigned to be 50 ohms so the input coupling loops were redesigned to be 50 ohms which is in the final design of the ETA. The tune loops in the output cavity resonator distorted the field distribution in the vicinity of the input coupling loops as was experienced with the input cavity resonator. The final ETA version of the output cavity resonator did not employ the electronic tuning for the same reasons mentioned earlier for the input cavity resonator. The mechanical tuning allows the output cavity resonator to be tuned over the frequency range of 336 MHz to 400 MHz.

The loaded Q was adjusted for 86 at 352 MHz and 76 at 400 MHz without the 12 power amplifier modules attached. During the testing phase, with the amplifiers connected, the capacitive output coupling probe was readjusted slightly to compensate for the amplifier loading. The loaded Q, Q_L, could not be measured conveniently after this adjustment was made. A measurement of the overall loaded Q which includes the input cavity resonator was measured during the final testing phase and these data appear in the test results.

2.3 Attenuators. The power limitations of available UHF PIN diodes prevents the practical use of a conventional PIN diode attenuator design, in which the diodes themselves dissipate all the power that must be absorbed. The circuit shown in Figure 2-17, however, permits two PIN diodes to control the absorption of input power by a resistor, while absorbing only a fraction of the power themselves. When the diodes are biased off, the full input power (minus the hybrid division insertion loss) is available at the output connector. When the control voltage biases the diodes into conduction, they absorb power and thus create a mismatch for the divider. The divider directs the resulting reflected power to its isolated port, where it is dissipated by the isolation resistor.

The amount of power reflected (and hence, the absorption) depends upon the effective resistance of the diodes, and is thus a function of the control voltage. Because of the balanced circuit configuration, the reflection coefficients at the two output ports are equal, and the input VSWR is therefore low and essentially independent of attenuation.

This technique was used in the ETA system for the input attenuator and for each of the individual amplifier balancing attenuators. Figure 2-18 is the schematic for the input attenuator while the 100 W amplifier attenuator schematic is shown in Figure 2-17.

The input attenuator was tested with 100 watts power input at 352, 368, 384, and 400 MHz. The attenuation ranged from approximately 0.8 to 14 dB. Maximum bias current required varied from 30 to 100 ma depending on frequency. Worst case VSWR was 1.2:1 and occurred at approximately 9.5 dB attenuation. When connected in the system the output power was able to be reduced from 1000 watts (at 368 MHz) down to 10.7 watts, a range of 19.7 dB. The apparent greater range is probably due to the system's non-linear gain characteristics, and would vary with different power levels and frequencies.

The attenuator was also tested at 150 watts input power. Some difficulty was experienced in adjusting the attenuator in the middle range of values. This was determined to be high power dissipation in the PIN diodes resulting in a varying value for diode resistance. At 200 watts input power the attenuator could not be adjusted except at high or low values where the diode dissipation is lowest.

The PIN diodes used are Unitrode UM4900 series and are rated at 37 watts dissipation. Two diodes in series (for best intermod) are used on each leg of the hybrid dividers. Unitrode has been consulted concerning this problem and has recommended using 3 diodes on each leg. However, time was not available to design and fabricate a new heat sink package for the input attenuator. By limiting the system input power to a level of 100 watts or less no problem is experienced.

The 100 watt amplifier attenuators were tested with 4 watts power input at 352, 368, 384, and 400 MHz. The attenuation ranged typically from 0.6 to 14 dB. Maximum bias current averaged about 20 ma. Worst case VSWR was 1.4:1.

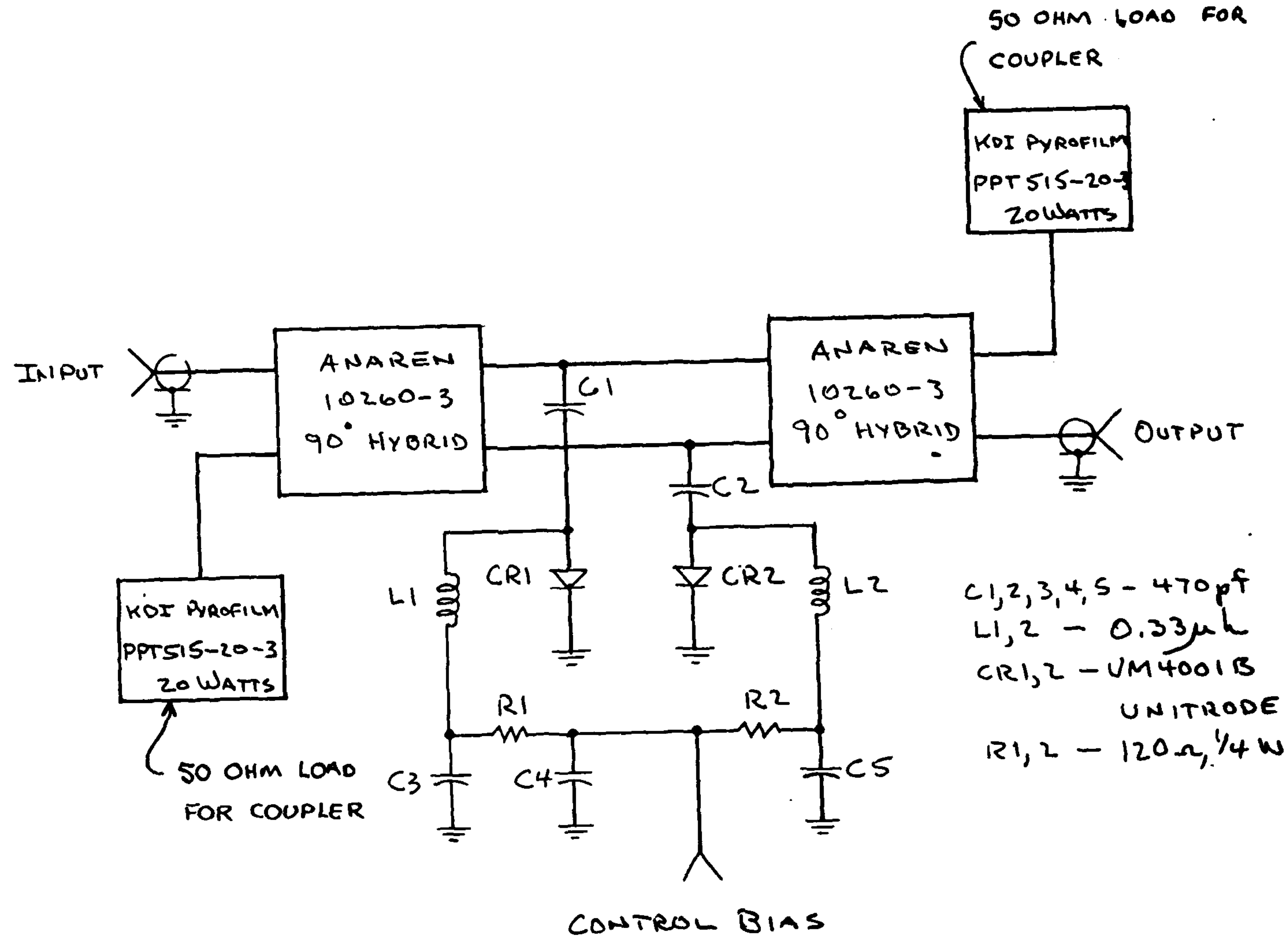


FIGURE 2-17 - ETA 100 W AMP INPUT ATTENUATOR

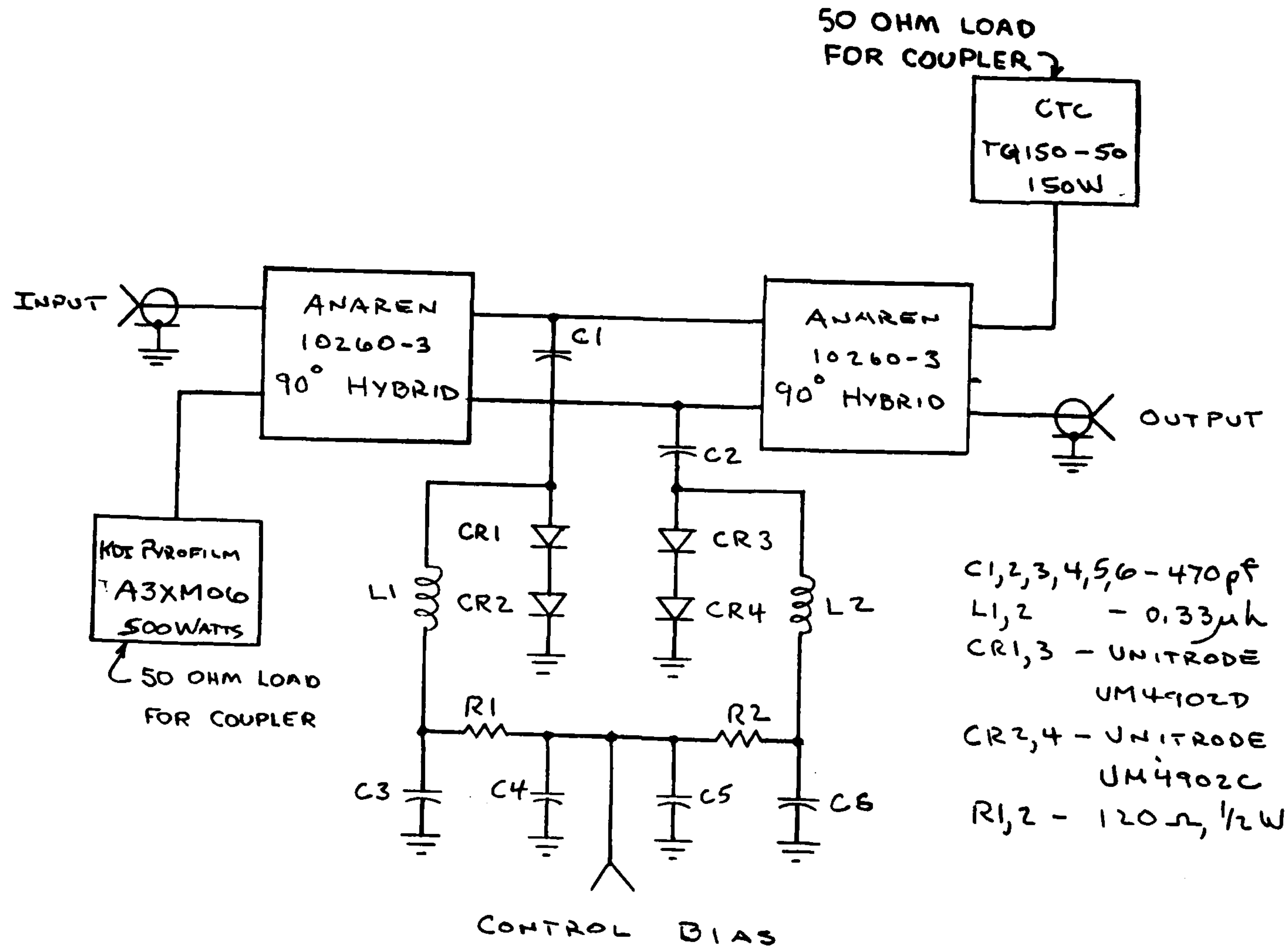


FIGURE 2-18 - ETA SYSTEM INPUT ATTENUATOR

The amplifier attenuators were included in the ETA system to provide a means for adjusting the balance between the twelve amplifiers. Since this was not possible, because of the interaction between amplifiers, these attenuators are not controlled. No bias is applied, thus all units are at minimum attenuation.

2.4 DC Switch. The purpose of the dc switch is to provide a means for rapidly disabling the rf power amplifiers in the event of a problem that would be detrimental to the reliability of the ETA system. There are 12, one for each amplifier. This function is accomplished by opening the path of the +28 Vdc to the transistors' collectors. Signals that will trip this function are: high VSWR (monitored at the system output), overvoltage (on the +28 Vdc source), and /or excessive current in the rf amplifier module.

Figure 2-19 is the schematic diagram of the dc switch. Operation is explained as follows: Q3 is the switching transistor. It is controlled by Q2 and Q5 and is either cut-off or saturated. Q2 is controlled by Q4 which together with Q5 forms a differential switch. The differential switch is driven by U2-3 and U2-6. A positive pulse "ON" signal at E6 resets the flip-flop and Q4, Q2, and Q3 are driven to saturation and the switch is "ON".

The total rf amplifier current passes through R1. The rf driver transistor current passes through R17. If either of these currents becomes too high Q1 or Q9 will switch "ON". This will cause a positive pulse to appear at U1-9, thus setting the R-S flip-flop. Q4 and Q2 will be cut-off and Q3 will lose its drive current. Q5 will saturate rapidly cutting off Q3. At the same time, Q6, Q7, and Q8 will saturate. These work like a crowbar circuit and discharge the energy storage capacitors on the rf amplifier down to 5 volts in less than 10 microseconds. R4 provides a means to adjust the total trip current to 10 amperes. Q9 trips at 1.8 to 1.9 amps and is not adjustable.

The dc switch is constructed on an aluminum finned heat sink. The rf amplifier variable attenuator is mounted on the same module. The "VSWR trip" function has been disabled in the final ETA system because of noise problems and false triggering. Also there is no "fail" signal since the micro-processor controller is not included.

2.5 Control System. The original concept for the control system used a microprocessor which sampled the rf output power level of each 100 watt amplifier module and made corresponding adjustment of each electronic input attenuator shown in Figure 2-17 to assure proper distribution of the total load power equally between the operational modules. As described in Section 2.2 the variation in cavity resonator input impedance varies unpredictably with frequency using electronic tuning and it became impractical to measure the power level at the output of each 100 watt module. Without the capability to measure power level, the concept of electronic power control becomes unworkable and the idea was abandoned. The design then reverted to a conventional manually controlled system through potentiometers and switches located on the front panel of the ETA chassis.

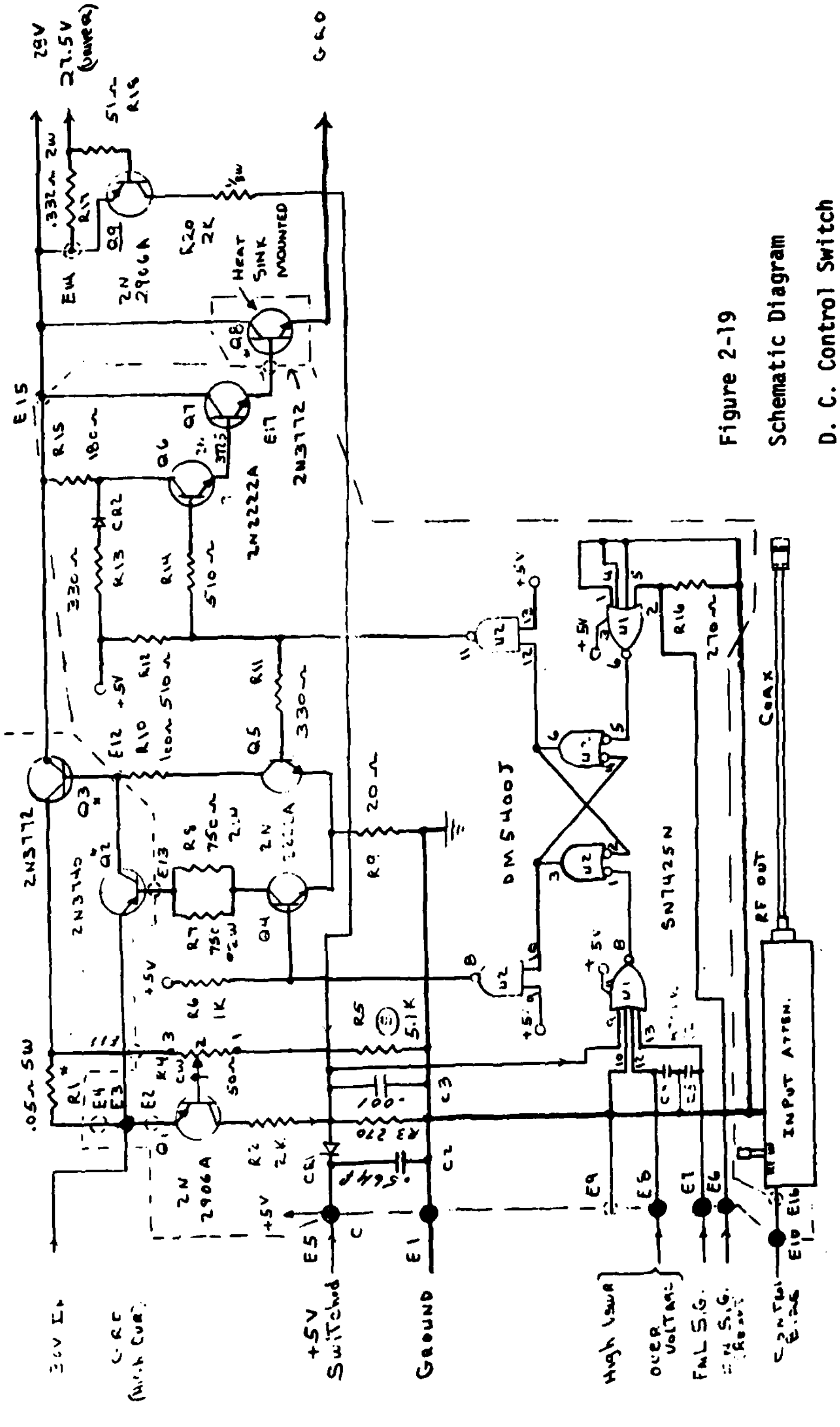


Figure 2-19

Schematic Diagram

D. C. Control Switch

2.6 Packaging and Cooling. The electronically tuned amplifier is designed to fit within the volume of a standard MS 91403-C2DATR chassis. The amplifier is comprised of the following module assemblies:

1 - Output Resonator

1 - Input Resonator

12 - 100 watt RF Modules

12 - Attenuator and dc Switch Modules

1 - Control Module

Figure 2-20 shows the entire ETA assembly with the cooling shroud attached. The cooling air enters at the left side of the assembly and flows over the cooling fins of the RF amplifier heat sinks and is exhausted at the right side of the assembly.

Figure 2-21 shows a cross section of the output resonator which has a 5" inside diameter and a 8-1/2" length. One of the 12 coupling loops for the RF amplifier outputs is shown. Other amplifiers are mounted around the circumference of the cylinder. Note that provision is made for adjustment of the coupling loop from outside the cavity. The loop is driven by a section of transmission line with a rotary joint at the point where the loop is attached at the resonator wall.

Figure 2-22 shows a longitudinal section through the output cavity. The connection to the output transistor is identified in the illustration. Note that each 100 watt module and its associated heat radiator can be removed from the main assembly for ease of maintenance. Note that the length of the center element of the cavity can be adjusted by means of an external knob in the base plate of the resonator. The input resonator (not shown) is located at the right of the output resonator and has a similar internal configuration. The resonators can be separated from each other (by removing the attaching bolts) for other mounting configurations. The dc switch module contains the "gain difference" attenuators (not used) with a 50 ohm connection to the coaxial cable which attaches to the input resonator coupling loops. The 28 Vdc power switch transistor is located on this module.

For ease of construction, elimination of sophisticated plating, and improved heat transfer Magnavox fabricated the experimental model ETA from copper. Eventual models of the device for airborne application would be fabricated of aluminum with plating. The experimental model is not pressurized; however airborne applications probably will require pressurization of the internal cavity area due to the high RF voltage present on the center conductor.

The weight of the composite system is 145 pounds.

2.6.1 Thermal Considerations. The cooling air is supplied by a Rotron fan (Model B-502) with a flow rate of 262 cfm at a pressure drop of 1.82" H_2O. The air flow is split among the 12 module sections and there is approximately 22 cfm per section.

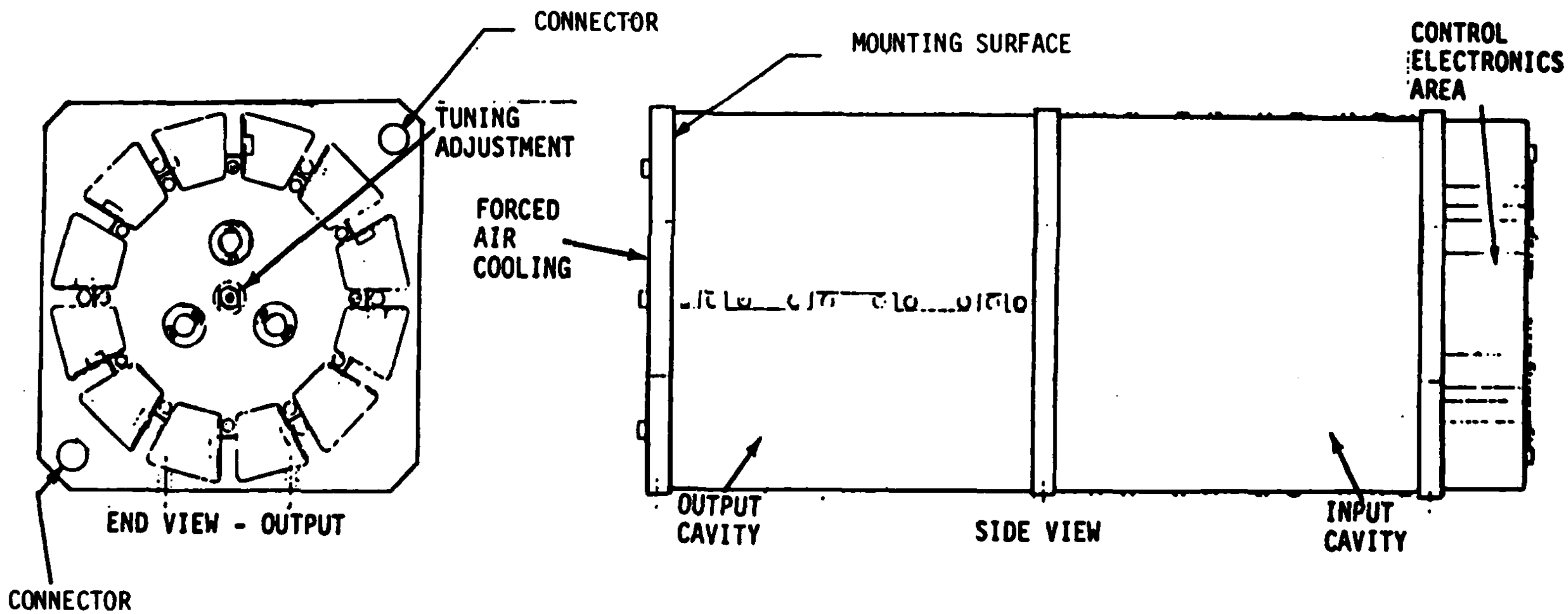


FIGURE 2-20. ETA INTERNAL OUTLINE

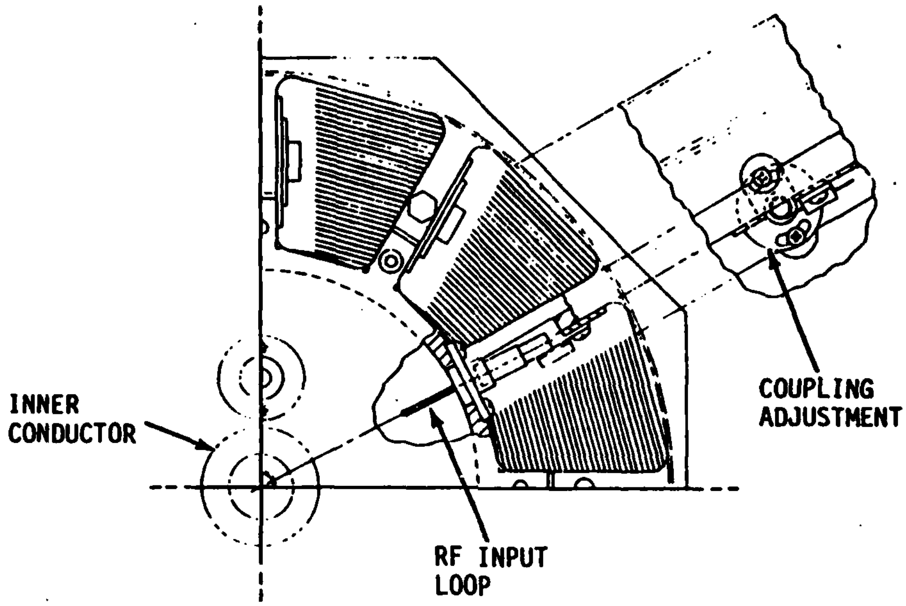


FIGURE 2-21. CROSS SECTIONAL VIEW - OUTPUT RESONATOR

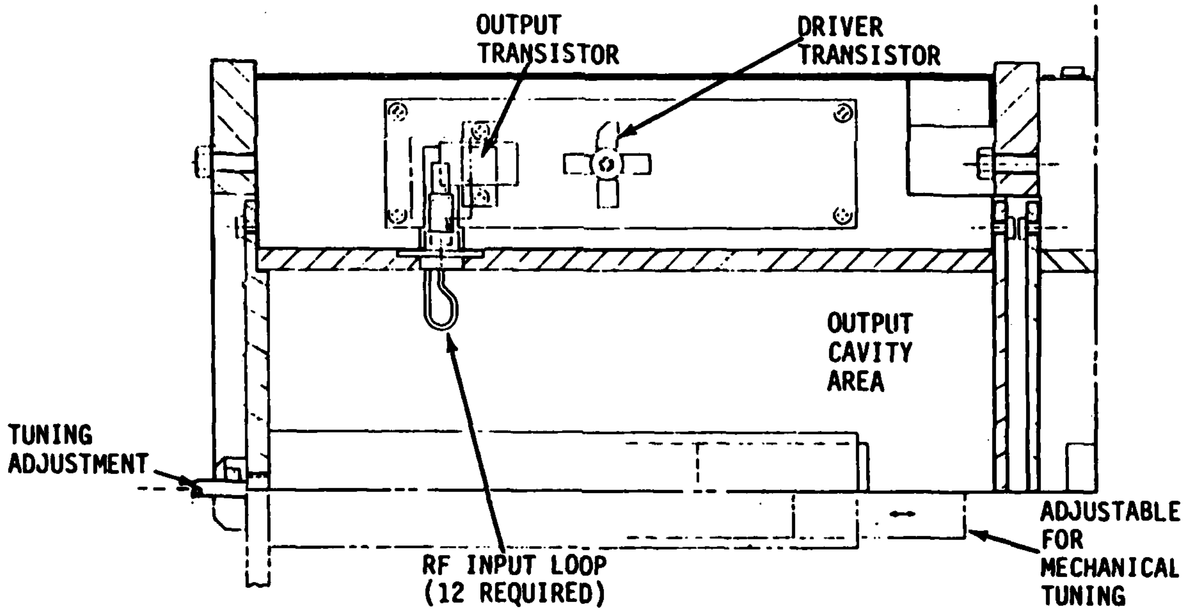


FIGURE 2-22. LONGITUDINAL SECTIONAL VIEW - OUTPUT RESONATOR

The C2M100-28A device on the RF module is the most critical component concerning thermal design. Due to the high wattage dissipation and the small mounting surface, the heat sink was fabricated from copper. The transmissibility of copper is twice that of aluminum. To provide optimum thermal paths, the device is mounted directly to the heat sink utilizing thermally conductive grease. The other high dissipating devices are mounted in the same manner. This will maintain a temperature differential of 10° C to the heat sink. The heat sink is 2" high x 3/8" thick x 8" long. To increase the effective cooling area, the rear of the heat sink has 20 full length fins which provides 352 square inches of cooling surface. This configuration maintains a temperature differential of 25°C.

The temperature differential from the heat sink to the cooling air is approximately 60°C. With the 60°C differential and the flow rate of 22 cfm per module the critical junction temperature of the devices can be held to a low termperature rise.

2.6.2 Module Accessibility. A top cover is used for access to the modules and components. The cover is equipped with RF gasketing to provide RF interference integrity. Power, control, and RF connectors are located on the front face along with an inlet air port.

3.0 TEST AND EVALUATION

3.1 Amplifier Performance. All amplifiers were tested into a 50 ohm, resistive load. Data was taken to permit characterizing each amplifier for gain, efficiency, and input return loss at each of the four frequencies of interest. This data is summarized in Table 3-1. Note that the efficiency is that of the output stage only. The driver current is included in the Table as a point of reference.

Amplifier performance is critical with regard to physical placement of some of the components. Reference is made to Figure 2-6 Schematic Diagram. The efficiency of the driver stage is highly dependent upon the placement of C12 and 13 and upon the values of C14 and 15. This is because of the "transparency" of Q2. The input return loss varies as the location of C10 and 11 is changed with respect to the edge of Q2, again due to the driver "transparency". The stray capacity of L13 to ground also has a significant effect on circuit performance.

It was found that the emitter leads of the output transistor must be shorted to the mounting flange right at the transistor body. This was necessary to prevent oscillation with a narrow band load such as a tunable filter or resonant cavity. No oscillation was experienced with a broad-band load. This problem was discussed in the section on amplifier design.

Before any amplifier was operated it was necessary to adjust the quiescent collector currents to standard values for class AB bias. 20 ma was selected for the driver and 30 ma for the output. The exact values are somewhat arbitrary and were based upon values used in another program. Adjustment was made by using potentiometers for R1 and R5 in the circuit and then replacing them with the nearest standard value carbon resistors. It was originally intended that the class of bias could be switched from A to AB to C. Tests on the breadboard amplifier indicated that Class C did not provide adequate gain. Class A operation was unstable without

TABLE 3-1

SN is Module Serial Number and not the same as Position Number
F is the test frequency.
G is the module gain.
RL is the amplifier input return loss
Io is the collector current of the output transistor.
eff is the rf output power divided by the dc input power of the output transistor [100/(28 x Io)] times 100.
I_D is the collector current of the driver transistor.

RF AMPLIFIER TEST RESULTS - 100 WATTS

SN	F	G	RL	Io	eff	I_D
	MHz	dB	dB	A	%	A
1	352	20.2	7.9	5.7	62.7	.0
	368	19.5	8.9	6.3	56.7	1.12
	384	19.3	9.5	6.4	55.9	1.2
	400	19.9	10.8	6.3	56.7	1.02
2	352	20.4	11.5	6.1	58.5	.86
	368	20.9	12.6	5.9	60.5	.94
	384	20.8	11.1	5.6	63.8	.98
	400	19.7	9.8	5.4	66.1	1.06
3	352	21.0	9.4	6.7	53.3	.75
	368	21.1	9.8	6.8	52.5	.82
	384	20.8	8.8	6.6	54.1	.92
	400	20.1	8.5	6.6	54.1	.84
4	352	19.6	10.0	6.1	58.5	.92
	368	20.3	10.5	6.0	59.5	.94
	384	20.6	10.3	5.6	63.8	.965
	400	20.1	9.4	5.4	66.1	.97
5	352	20.4	8.9	7.2	49.6	.7
	368	21.0	9.4	6.9	51.8	.73
	384	21.1	8.6	6.5	54.9	.75
	400	20.4	8.2	6.4	55.8	.72
6	352	19.9	8.7	6.3	56.7	.9
	368	20.2	9.4	6.4	55.8	1.0
	384	20.0	9.6	6.4	55.8	1.1
	400	19.7	9.1	6.4	55.8	.96

SN	F	G	RL	Io	eff	I_D
	MHz	dB	dB	A	%	A
7	352	19.9	8.5	6.6	54.1	.8
	368	20.4	8.9	6.8	52.5	.88
	384	19.5	8.9	6.4	55.8	1.1
	400	18.5	9.3	6.5	54.9	1.0
8	352	20.7	10.7	6.5	54.9	.8
	368	20.8	10.6	6.5	54.9	.88
	384	20.2	10.2	6.4	55.8	1.05
	400	19.1	9.8	6.6	54.1	1.06
9	352	20.3	8.0	6.9	51.8	.81
	368	20.3	8.6	7.0	51.0	.96
	384	20.4	9.2	6.6	54.1	1.1
	400	19.5	9.3	6.6	54.1	1.1
10	352	19.8	11.5	5.8	61.6	1.02
	368	20.2	11.8	5.8	61.6	1.12
	384	20.0	10.4	5.6	63.8	1.22
	400	19.7	9.5	5.5	64.9	1.02
11	352	20.4	11.0	6.1	58.5	.96
	368	20.8	11.9	5.9	60.5	.98
	384	20.9	11.1	5.5	64.9	.94
	400	19.9	10.4	5.4	66.1	1.0
12	352	19.9	9.0	7.4	48.2	.84
	368	20.9	11.0	6.9	51.7	.86
	384	21.3	9.5	6.2	57.6	.93
	400	20.8	9.5	6.2	57.6	.78

extensive control circuits. It was found that the operational parameters of the rf transistors vary widely with bias and would necessitate revision of the rf coupling circuitry for each class of bias. Since revision of the rf circuitry is not feasible, operation with Class AB bias was incorporated in the amplifier designs.

3.2 Attenuator Performance. The input attenuator was tested with 100 watts power input at 352, 368, 384, and 400 MHz. The attenuation ranged from approximately 0.8 to 14 dB. Maximum bias current required varied from 30 to 100 ma. depending on frequency. Worst case VSWR was 1.2:1 and occurred at approximately 9.5 dB attenuation. Additional tests were run as low as 5 watts and as high as 150 watts input power. Figure 3-1 shows the upper and lower limits for all conditions of input power, attenuation and frequency. A problem with high power dissipation in the diodes is discussed in the section on attenuator design.

The amplifier attenuators were tested with 4 watts power input at 352, 368, 384, and 400 MHz. The attenuation ranged typically from 0.6 to 14 dB. Maximum bias current averaged about 20 ma. Worst case VSWR was 1.4:1. Figure 3-2 shows the upper and lower limits for all conditions of attenuation and frequency and is applicable to all 12 attenuators.

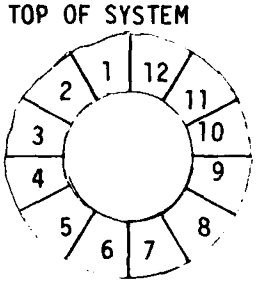
3.3 DC Switch Performance. All switches were tested with normal and maximum load currents. The switch was adjusted to trip at 10 amperes total amplifier module current. A driver amplifier current between 1.8 and 1.9 amperes tripped out the switch. With a 7 ampere load the output voltage to the output transistor varied between 28.57 and 28.86 Vdc. A 1 ampere load provided a voltage between 28.27 and 28.56 to the driver transistor. A fail signal caused the output voltage to reduce to 5 volts in 9.0 to 14.0 microseconds at no load and 7.0 to 8.5 microseconds at full load.

One problem was experienced with the dc switches during system operation. Occasionally a switch would trip out when the system was operating well below its maximum limits. This was due to electrical transients in equipment operating in the vicinity of the ETA. Capacitors C4 and C5 (see schematic diagram, figure 2-19) were added to reduce this sensitivity but did not completely solve the problem.

3.4 ETA System Test. All data in this section and the sections following, unless otherwise noted, was taken with the system outside its case and the outer cowling removed from the resonator assembly.

The positions of the amplifiers and switch modules are defined in the following diagram:

TOP OF SYSTEM



Viewed From the Panel End

POSITION	AMP. SN	SW. SN
1	10	2
2	4	10
3	7	9
4	9	8
5	8	7
6	6	11
7	2	5
8	5	4
9	3	3
10	11	6
11	12	1
12	1	12

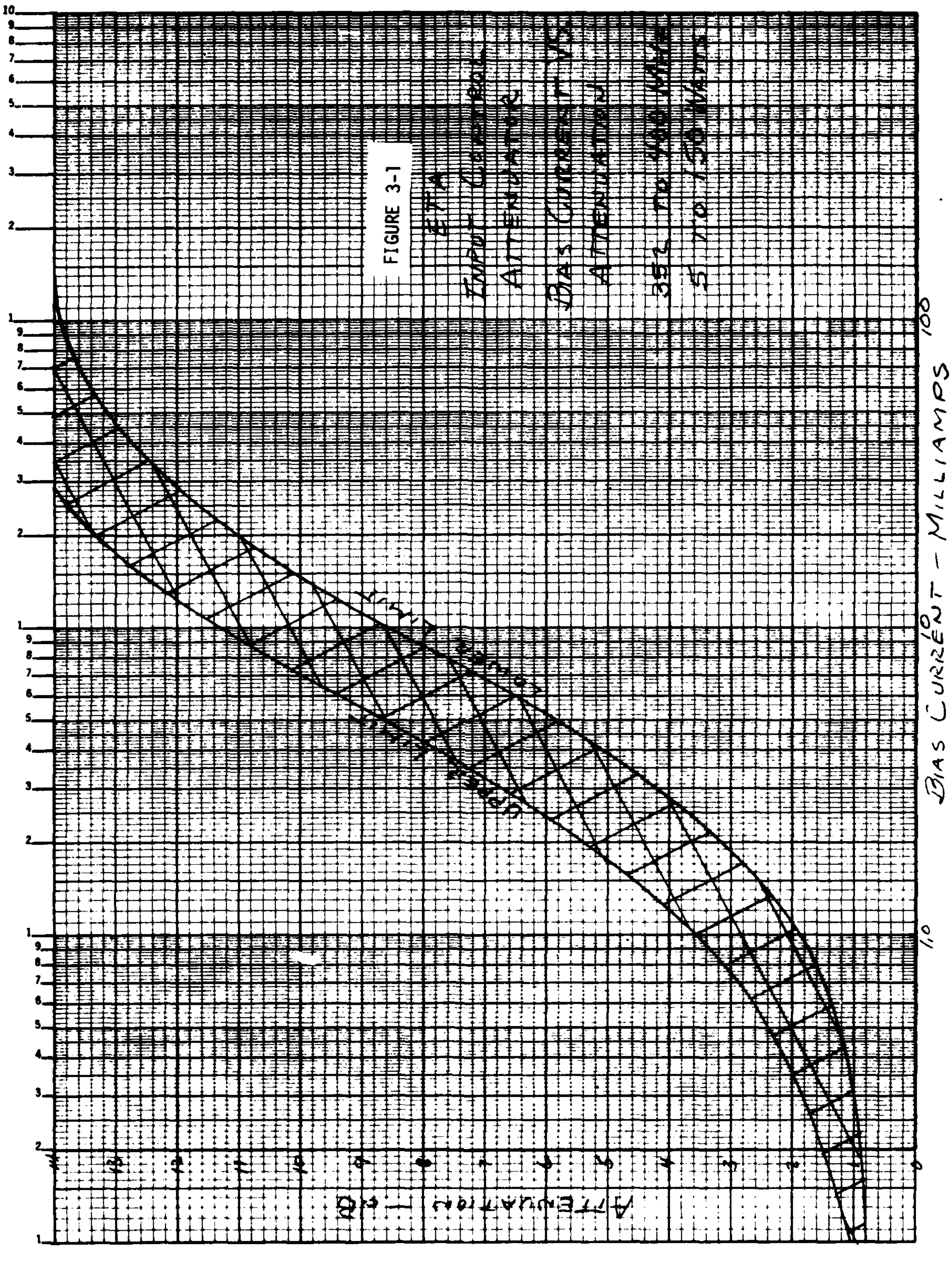


FIGURE 3-1
ETA
INPUT CONTROL
ATTENUATOR
BIAS CURRENT VS.
ATTENUATION
35L TO 400 MHz
5 TO 150 WATTS
BIAS CURRENT - MILLIAMPS
ATTENUATION — dB
LIMIT
UPPER LIMIT

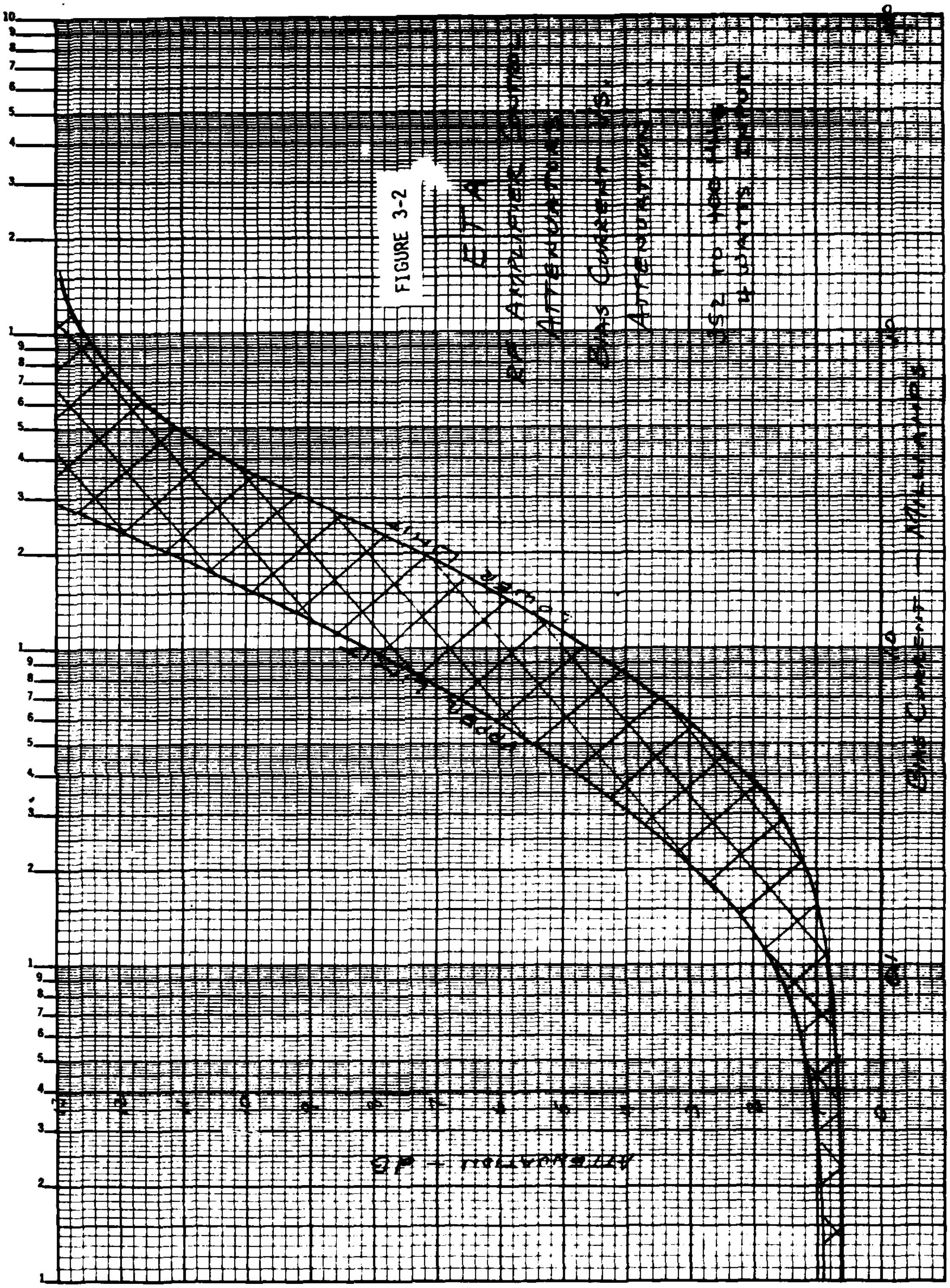


FIGURE 3-2

The resonator input and output probe connectors are in the center of position #11. The 100 watt amplifier input and output coupling loops are arranged axially with respect to the cavity except for position #7 which has the amplifier coupling loop rotated approximately 60° from the normal position. This was necessary to obtain the best balance on the dc current for that amplifier (SN-2) at all frequencies.

3.4.1 ETA System Test - CW Unmodulated. The ETA system was tested for a 1KW CW output at 352, 368, 384, and 400 MHz. Figure 3-3 illustrates the test set-up. The Marconi 1066B/1 and Avantek GPD, 401 were not used in these tests. The test data is shown in Table 3-2. All power readings shown are corrected for coupler and attenuator loss variations. Maximums, minimums, and total currents are indicated. Each reading is the sum of the output, driver and quiescent dc switch current. All system attenuators were operated with zero bias (minimum attenuation).

The dB levels for the second harmonics shown were directly from the spectrum analyzer. When a tunable filter was used to trap out the fundamental signal at 368 MHz the second harmonic measured -98dBc.

A thermocouple was connected to the copper heat sink adjacent to the output transistor on the 100 watt amplifier module in position #8. After 15 minutes of operation at 1KW and 400 MHz the indicated temperature had stabilized at 40.2°C.

When the original test data was taken some difficulty was experienced with the power meter for the input reflected power. Table 3-3 contains data taken in a subsequent test for return loss. Since the data was taken with all internal attenuators at zero bias (minimum attenuation) the results represent worst case.

After completion of all system tests the ETA was assembled into its final package. A control was added to the front panel to permit adjustment of the system input attenuator. Table 3-4 shows the data of a test run to compare amplifier currents. Table 3-5 shows the complete data when the input power is held at 25 watts and the output is controlled to 800 watts by the attenuator control.

3.4.2 ETA Amplitude Modulation Test. The ETA system was tested for its reponse to amplitude modulation. Figure 3-3 is the block diagram showing the equipment interconnection. Tests were run using 1 KHz modulation at 30 and 90% with a power output of 200 watts average. Figures 3-4 and 3-5 show the typical responses with the upper traces being the input signals and the lower traces the output. These waveforms were viewed on an oscilloscope, the signals being obtained from the 20 KHz IF output of an HP8405A vector voltmeter.

Table 3-6 contains the distortion measurements as read from an HP334A distortion analyzer. Notice that at 400 MHz the input distortion is greater than the output distortion. This is apparently due to the system non-linearities cancelling some of the input distortion components.

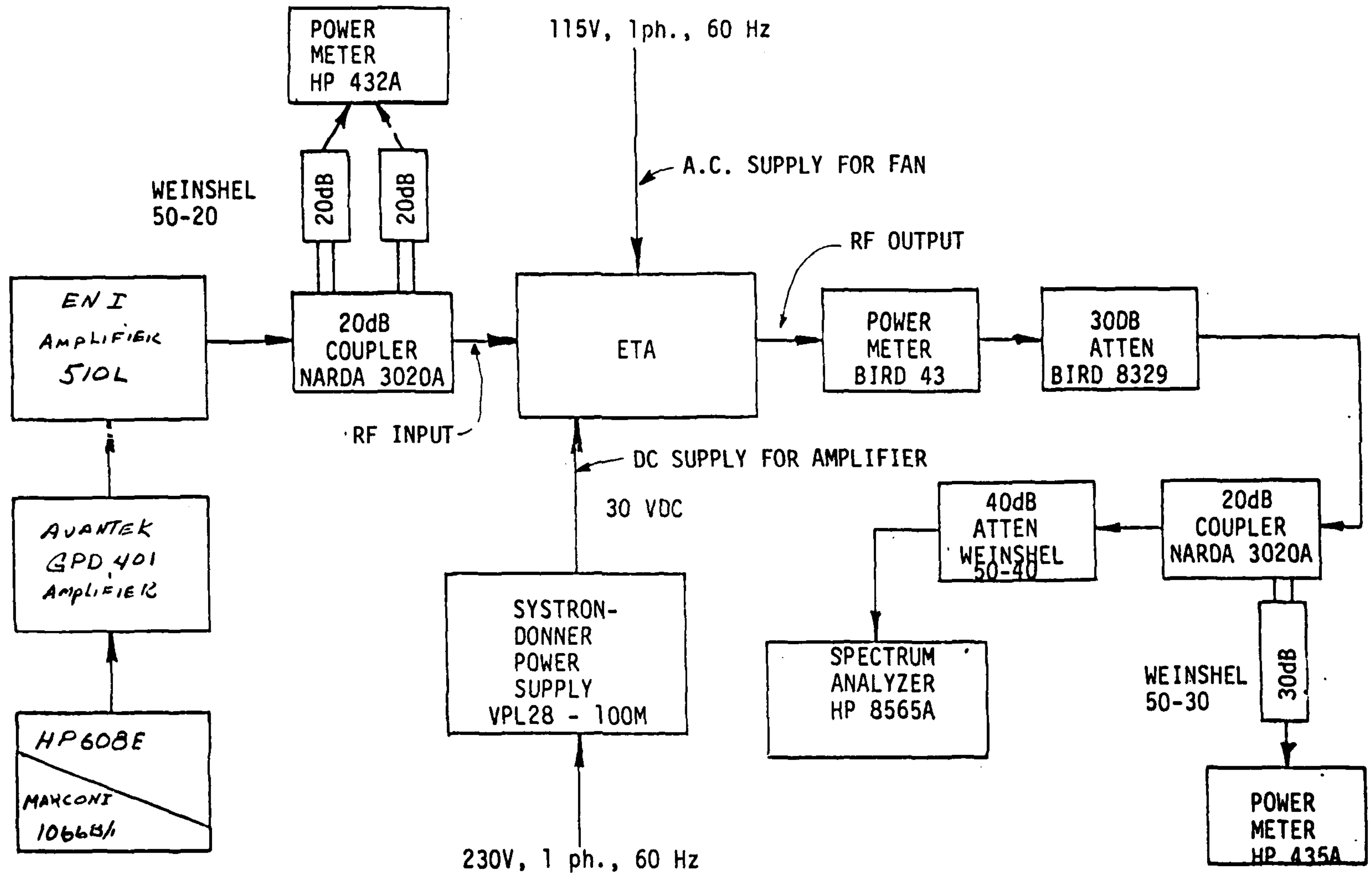


FIGURE 3-3 BLOCK DIAGRAM, ETA TEST SET-UP

TABLE 3-2

CW TEST 1KW NO MODULATION

FREQ.	POWER OUTPUT	POWER INPUT	GAIN	IN. REF. POWER*	RETURN LOSS	COLLECTOR VOLTAGE	TOTAL CURRENT	MAXIMUM CURRENT		MINIMUM CURRENT		EFFICIENCY
MHz	WATTS	WATTS	dB	WATTS	dB	V	A	Pos.#	A	Pos.#	A	%
352	958	9.49	20.0	5.75	2.2	28.21	79.55	4	7.6	6	6.05	42.7
368	989	7.48	21.2	.45	12.2	28.58	68.5	4	6.3	7	4.8	50.5
384	997.6	8.77	20.6	.86	10.1	28.59	67.5	3	6.7	6	4.3	51.7
400	985.5	13.4	18.7	2.37	7.5	28.29	73.9	3,8	7.2	6	4.9	47.1

READINGS VARIABLE AND NON-REPEATABLE

POSITION #7 OUTPUT LOOP IS ROTATED APPROXIMATELY 60° FROM NORMAL POSITION

TEMPERATURE AT POSITION #8 WAS 40.2°C AFTER 15 MINUTES AT 1 KW (STABILIZED)

*INPUT REFLECTED POWER = POWER DEVELOPED BY THE SOURCE AND NOT DISSIPATED IN THE LOAD DUE TO INPUT MISMATCH.

TABLE 3-3

CW TEST 1 KW NO MODULATION

FREQ	POWER INPUT	IN. REF. POWER*	RETURN LOSS
MHz	WATTS	WATTS	dB
352	11.4	.59	12.9
368	9.4	.235	16.0
384	10.0	.80	11.0
400	10.5	2.13	6.9

TABLE 3-4

CW TEST 1 KW NO MODULATION

FREQ	POWER OUTPUT	TOTAL CURRENT	MAXIMUM CURRENT		MINIMUM CURRENT	
MHz	WATTS	A	Pos.#	A	Pos.#	A
352	967	78.75	4	8.0	2,6	6.0
368	963	68.6	10	6.4	7	5.0
384	955	66.5	3	6.6	6	4.3
400	946	72.2	3	7.2	6	4.8

ETA SYSTEM ASSEMBLED INTO ITS FINAL PACKAGE

*INPUT REFLECTED POWER = POWER DEVELOPED BY THE SOURCE AND NOT DISSIPATED IN THE LOAD DUE TO INPUT MISMATCH.

TABLE 3-5

CW TEST 800W NO MODULATION

ETA SYSTEM ASSEMBLED INTO ITS FINAL PACKAGE

FREQ.	POWER OUTPUT	POWER INPUT	GAIN	IN. REF. POWER*	RETURN LOSS	COLLECTOR VOLTAGE	TOTAL CURRENT	MAXIMUM CURRENT		MINIMUM CURRENT	
MHz	WATTS	WATTS	dB	WATTS	dB	V	A	Pos.#	A	Pos.#	A
352	800	25	15.1	.075	25.2	---	68.65	4	6.65	6	5.2
Max. Attenuation Pout=60 W, R.L. = 17.8 dB; Min. Attenuation Pi = 8.5W, Pout=800W, G = 19.7dB											
368	800	25	15.1	.12	23.2	---	57.3	10	5.4	6	4.1
Max. Attenuation Pout=100 W, R.L. = 18 dB; Min. Attenuation Pi = 6.1W, Pout=800W, G = 21.2 dB											
384	800	25	15.1	.12	23.2	---	57.0	3	5.7	6	3.6
Max. Attenuation Pout=90 W, R. L. = 19.5 dB; Min. Attenuation Pi = 7.3W, Pout=800W, G = 20.4 dB											
400	800	25	15.1	.81	14.9	---	61.8	3	6.3	6	3.9
Max. Attenuation Pout=60 W, R.L. = 22.8 dB; Min. Attenuation Pi = 10W, Pout=800 W, G = 19.0 dB											

*INPUT REFLECTED POWER = POWER DEVELOPED BY THE SOURCE AND NOT DISSIPATED IN THE LOAD DUE TO INPUT MISMATCH.

3.4.3 ETA Frequency Modulation Test. The ETA system was also tested for its response to frequency modulation. Figure 3-3 is also applicable to this test. A 1 KHz modulation with an 8 KHz deviation was used at a power output of 800 watts average. Figures 3-6 and 3-7 illustrate the typical input and output responses as seen on the spectrum analyzer. The results were satisfactory. Table 3-7 contains the data.

3.4.4 ETA System Back Intermodulation Test. The purpose of the back intermodulation test is to determine the interference generated as a result of two co-located high power transmitters. This condition was simulated for the ETA system by connecting it as shown in Figure 3-8.

The ratio of the received power to the radiated power for antennas operating at a wavelength λ and at separation d in free space is:

$$\frac{P_2}{P_1} = \left(\frac{\lambda}{4\pi d}\right)^2 G_1(\theta, \phi)\, G_2(\theta, \phi)$$

G_1 and G_2 are equal to one for isotropic radiators. If an antenna spacing of three feet is assumed at 400 MHz then:

$$\frac{P_2}{P_1} = \left(\frac{29.53}{4\pi\, 36}\right)^2 = 4.26 \times 10^{-3}$$

This is 4.26 watts at the ETA output antenna for a 1 KW co-located transmitter, or 0.426 watts for a 100 watt transmitter. Allowing for a 0.25 dB cable loss, the powers would be 4 and 0.4 watts. These were the numbers used for this test. The ETA output was limited to approximately 800 watts to limit the stress on the output transistors.

The test results are shown in Figures 3-9 through 3-16. They show that the intermodulation varies between 10 and 12.7 dB below the interfering signal where it is one MHZ below the ETA center frequency. Notice that there are only slight differences in results between 4 watts and 0.4 watts interfering signal. These results must be viewed with caution however, since Magnavox did not have available high Q, narrow-band notch filters to remove the two strong signals from the spectrum analyzer input. Figure 3-17 is representative of additional tests run with the two signals spaced greater than one MHz apart (in this instance, 9 MHz). Notice that this intermodulation is lost in the noise floor.

3.4.5 System Test of Input Attenuator. For reasons of safety, the ETA system was tested with no bias on the input attenuator (i.e., minimum attenuation). Two tests were performed to demonstrate the functioning of the attenuator. In the first, the system was operated at 1 KW output power while the control bias voltage (external source) was increased. It was possible to reduce the power output to 10.7 watts, an adjustment range of 19.7dB. In the second test a fixed power input level of 37 watts was applied to the ETA. The output level could be reduced only to 200 watts under these conditions, since part of the attenuator range was used in reducing the 37 watts to the required operating level. The test data is shown in Table 3-8.

TABLE 3-6 DISTORTION TESTS

AM TEST 1 KHz MODULATION
200 WATTS OUTPUT POWER

FREQ.	PERCENT MODULATION	PERCENT DISTORTION	PERCENT DISTORTION
MHz		INPUT	OUTPUT
352	30	1.34	1.375
352	90	2.45	4.62
368	30	1.15	1.20
368	90	2.40	7.50
384	30	1.25	1.15
384	90	3.5	5.70
400	30	1.65	1.37
400	90	10.10	8.30

FIGURE 3-4

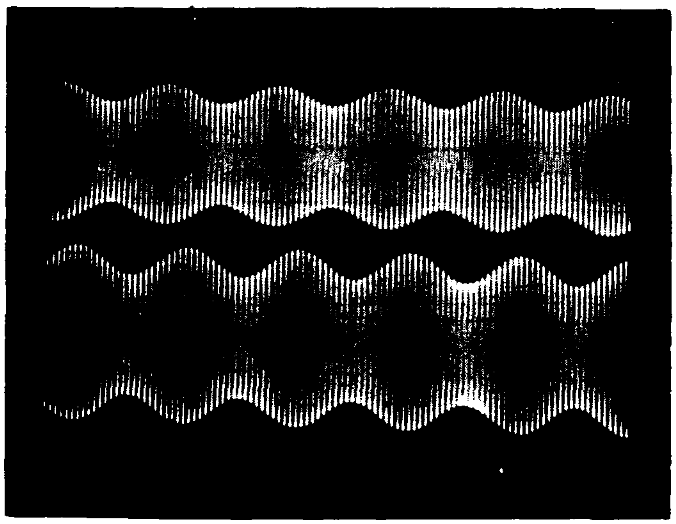


FIGURE 3-5

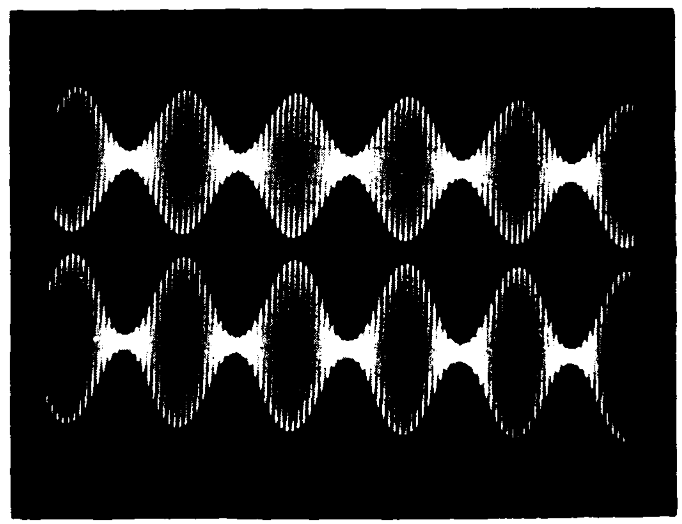


TABLE 3-7 FM TEST DATA

FM TEST 1 KHz MODULATION, 8 KHz DEVIATION

FREQ.	POWER OUTPUT	POWER INPUT	GAIN	IN.REF. POWER*	RETURN LOSS	COLLECTOR VOLTAGE	TOTAL CURRENT	MAXIMUM CURRENT		MINIMUM CURRENT		EFFICIENCY
MHz	WATTS	WATTS	dB	WATTS	dB	V	A	Mod.#	A	Mod.#	A	%
352	801	6.88	20.7	.096	18.6	28.73	70.0	4,10	6.6	5,6	5.4	30.8
368	800	5.33	21.8	.043	20.9	28.89	59.2	4,10	5.5	6	4.2	46.8
384	800	6.28	21.1	.146	16.3	28.86	58.5	3	5.8	6	3.6	47.4
400	797	9.81	19.1	1.84	7.3	28.77	64.8	3,8	6.4	6	4.1	42.8

READINGS VARIABLE AND NON-REPEATABLE

*INPUT REFLECTED POWER = POWER DEVELOPED BY THE SOURCE AND NOT DISSIPATED IN THE LOAD DUE TO INPUT MISMATCH.

FIGURE 3-6
INPUT

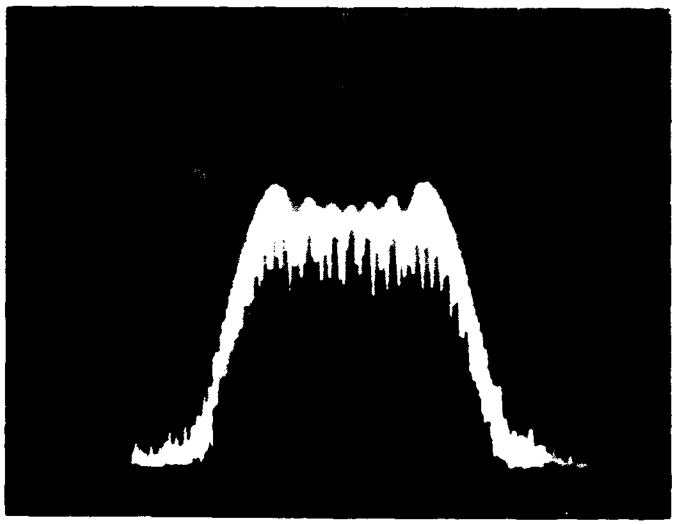
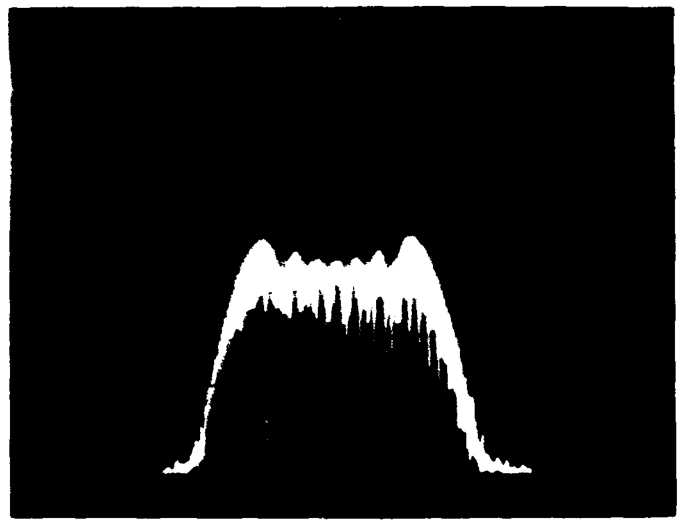


FIGURE 3-7
OUTPUT



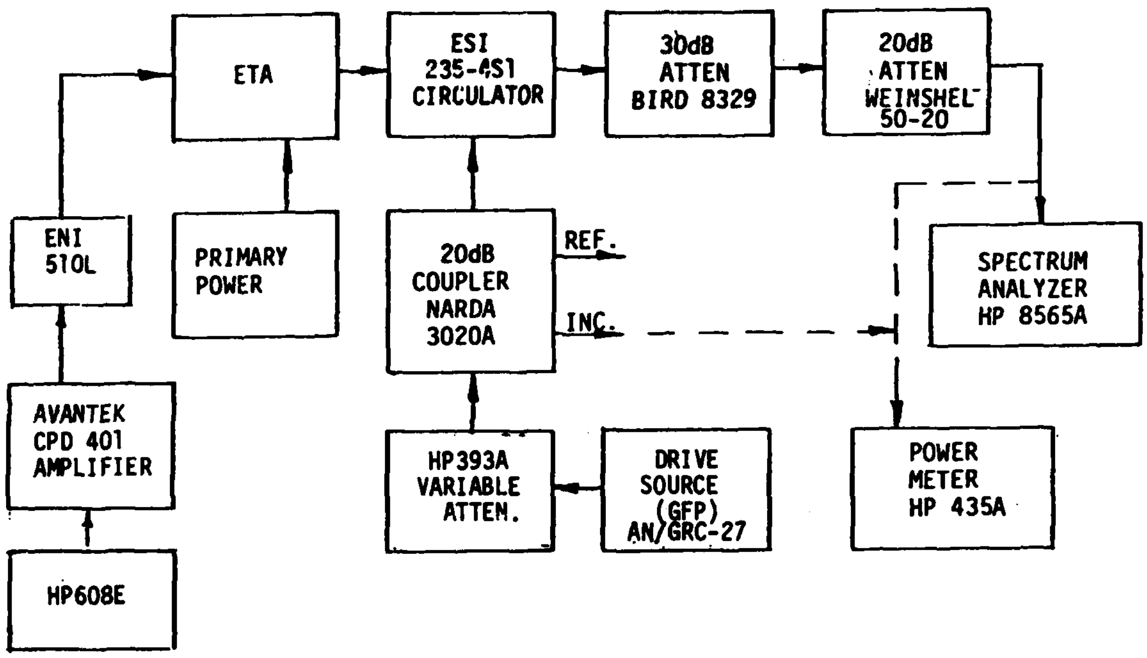


FIGURE 3-8 - BLOCK DIAGRAM, ETA BACK INTERMODULATION TEST

BACK INTERMODULATION TEST - 352 MHz

Po = 800 watts
Interfering Signal
= 0.4 watts @ 351 MHz

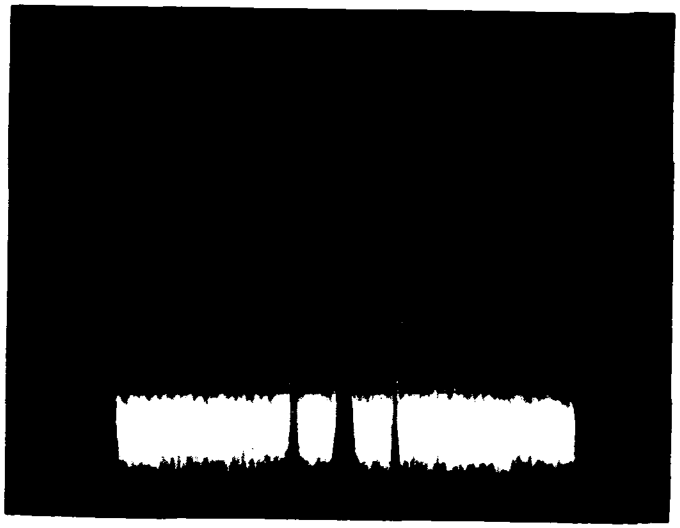


FIGURE 3-9

Po = 800 watts
Interfering Signal
= 4.0 watts

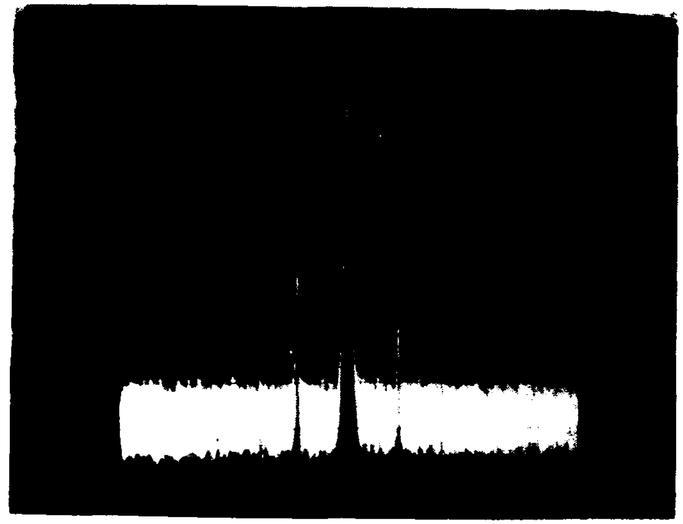


FIGURE 3-10

BACK INTERMODULATION TEST - 368 MHz

Po = 800 watts
Interfering Signal
= 0.4 watts @ 367 MHz

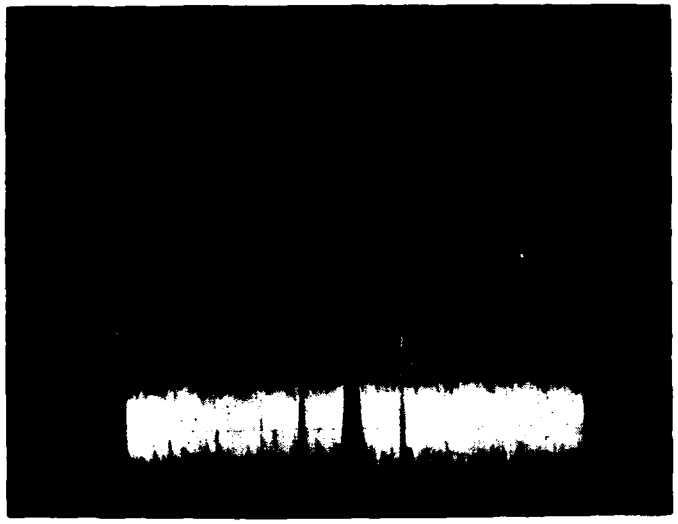


FIGURE 3-11

Po = 800 watts
Interfering Signal
= 4.0 watts @ 367 MHz

FIGURE 3-12

BACK INTERMODULATION TEST - 384 MHz

Po = 800 watts
Interfering Signal
= 0.4 watts @ 383 MHz

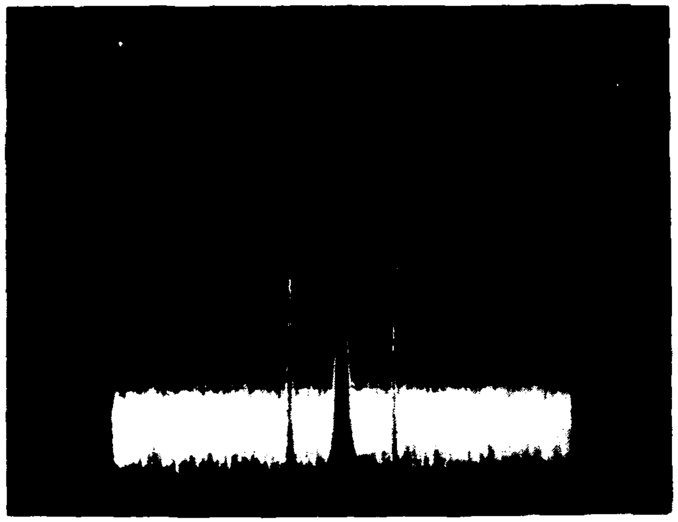


FIGURE 3-13

Po = 800 watts
Interfering Signal
= 4.0 watts @ 383 MHz

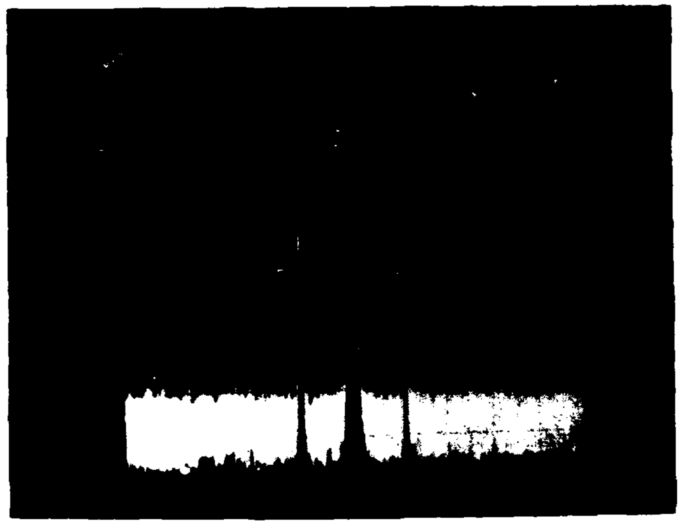


FIGURE 3-14

BACK INTERMODULATION TEST - 400 MHz

Po = 740 watts
Interfering Signal
= 0.4 watts @ 399

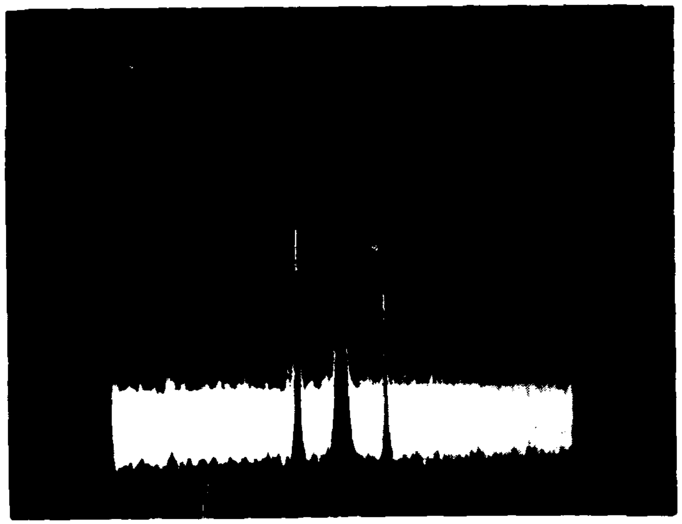


FIGURE 3-15

Po = 750 watts
Interfering Signal
= 4.0 watts @ 399 MHz

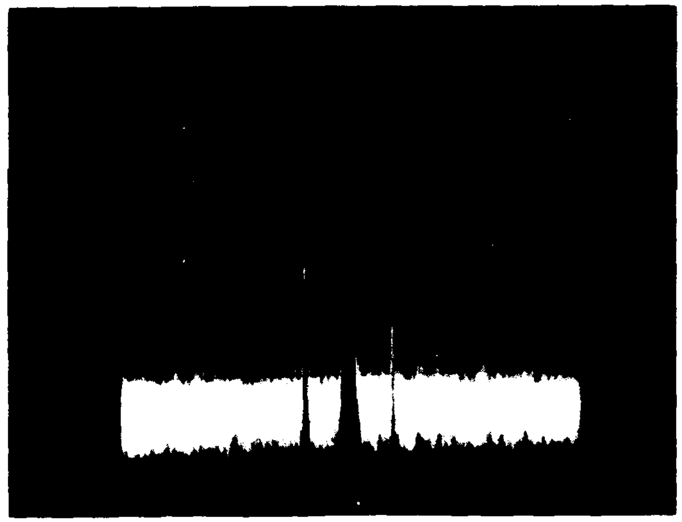


FIGURE 3-16

BACK INTERMODULATION TEST - 377 MHz

Po = 800 watts
Interfering Signal
= 0.4 watts @ 377 MHz

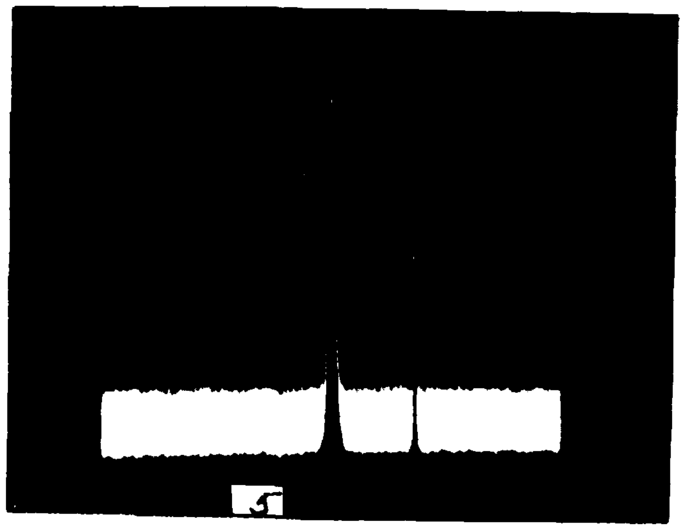


FIGURE 3-17

TABLE 3-8

SYSTEM TEST OF INPUT ATTENUATOR (368 MHz)

POWER OUTPUT	ATTENUATION	BIAS VOLTAGE
WATTS	dB	VOLTS
1000*	0	0
794	1	+.91
631	2	+1.02
501	3	+1.10
251	6	+1.33
126	9	+1.55
100	10	+1.63
31.6	15	+2.23
10.7	19.7	+7.21

*2nd Harmonic IS 98 dB below the fundamental.

At 400 MHz. 37 watts input power, output could be reduced to 200 W.

3.4.6 System Bandwidth. The 3 dB bandwidth of the ETA system was measured for a center frequency power output of 100 watts. For each frequency the input power was held constant and the frequency varied up and down until the output power dropped to 50 watts. Each frequency was then recorded and the bandwidth calculated. Note that these results are applicable to the system as a whole and do not indicate the bandwidth of the output resonator. The results indicate an approximate bandwidth of 3.25 MHz at 352 and 400 MHz, and 2.8 MHz at 368 and 384 MHz. The data is shown in Table 3-9.

4.0 CONCLUSIONS AND RECOMMENDATIONS

4.1 R.F. Power Summing. The investigation of the ETA system has shown that resonant cavities can be used for efficient rf power summing when tuning is accomplished through adjustment of the length of the coaxial inner element. Summing with electronic tuning was not efficient. Unless special attention is given to amplitude and phase balance of individual amplifier modules, more than 6 summing inputs to a manually tuned cavity is not recommended. A better approach than that explored under the contract would be external summing at moderate power levels (100 to 200 watts) using 3 dB hybrid couplers (produces well matched 50 ohm amplifier loads) and summing of hybrid outputs in the resonator to secure the final high power output.

4.2 Electronic Tuning. The technique for electronically tuning a resonant cavity described herein was successfully demonstrated. It was possible to shift the frequency 3 MHz at 350 watts. The shortcomings of electronic tuning are limited power and frequency range. Also, the distortion of the RF field within the cavity precludes tuning and summing at the same time using this particular technique. It is recommended that investigations continue in the area of using more tuning loops, selection of PIN diodes for optimum breakdown voltage and possible techniques for electronically changing the length of the cavity center conductor.

4.3 Impedance Match to Cavity. Experiments proved that very large coupling loops were required for rf amplifiers with a low output impedance. The more amplifiers to be summed, the larger the required loops. When the amplifiers were modified to operate into 50 ohms, the required loops were of a reasonable size. The biggest problem is the lack of isolation between amplifiers. If automatic leveling is to be used, isolation is a must. Magnavox recommends using 90° hybrid couplers and connecting the amplifiers in 6 pairs as the next logical step.

4.4 Amplifiers. The rf amplifiers, as designed, are performing adequately in the ETA system. It is recommended that, for future work, the design be improved to the point that "tweaking" is not required when a transistor is changed or a new amplifier built. Reliability could be improved by not operating each device at its maximum rating. A survey of new devices, such as the Motorola XRF 328, should be made.

4.5 Attenuators. The rf input attenuator was demonstrated to operate satisfactorily at rf input powers up to 150 watts. It is recommended that additional work be done, such as increasing the number of diodes used, to permit operation at powers as high as 200 watts. The heat sink design should be modified to increase the air flow. Other circuits should be investigated for a more fail-safe design.

TABLE 3-9

SYSTEM 3 dB BANDWIDTH
(Constant Input Power)

			Pout - 3dB		
fc MHz	Pin W	Pout W	fL MHz	fH MHz	f MHz
352	1.27	100	350.23	353.5	3.27
368	0.97	100	366.93	369.7	2.77
384	1.09	100	382.39	385.23	2.84
400	1.70	100	398.37	401.61	3.24

4.6 Control System. A microprocessor control system was designed for the ETA system to control the frequency and power. Since electronic tuning and power balancing could not be incorporated into the final system, his controller was not implemented. It is recommended that, as the problems are solved, the controller be planned for in the resulting systems.

4.7 Performance Limitations. The ETA system was successfully operated at powers up to 1000 watts over the frequency range of 352 to 400 MHz with both unmodulated CW and FM. 200 watts of average rf power was obtained with 100% AM (1 KHz modulation). Operation with input power levels as high as 34 watts was demonstrated using the internal attenuator. It is possible to operate with higher input power but it is recommended that it be limited to 25 watts for system protection.